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Microwaves & RF

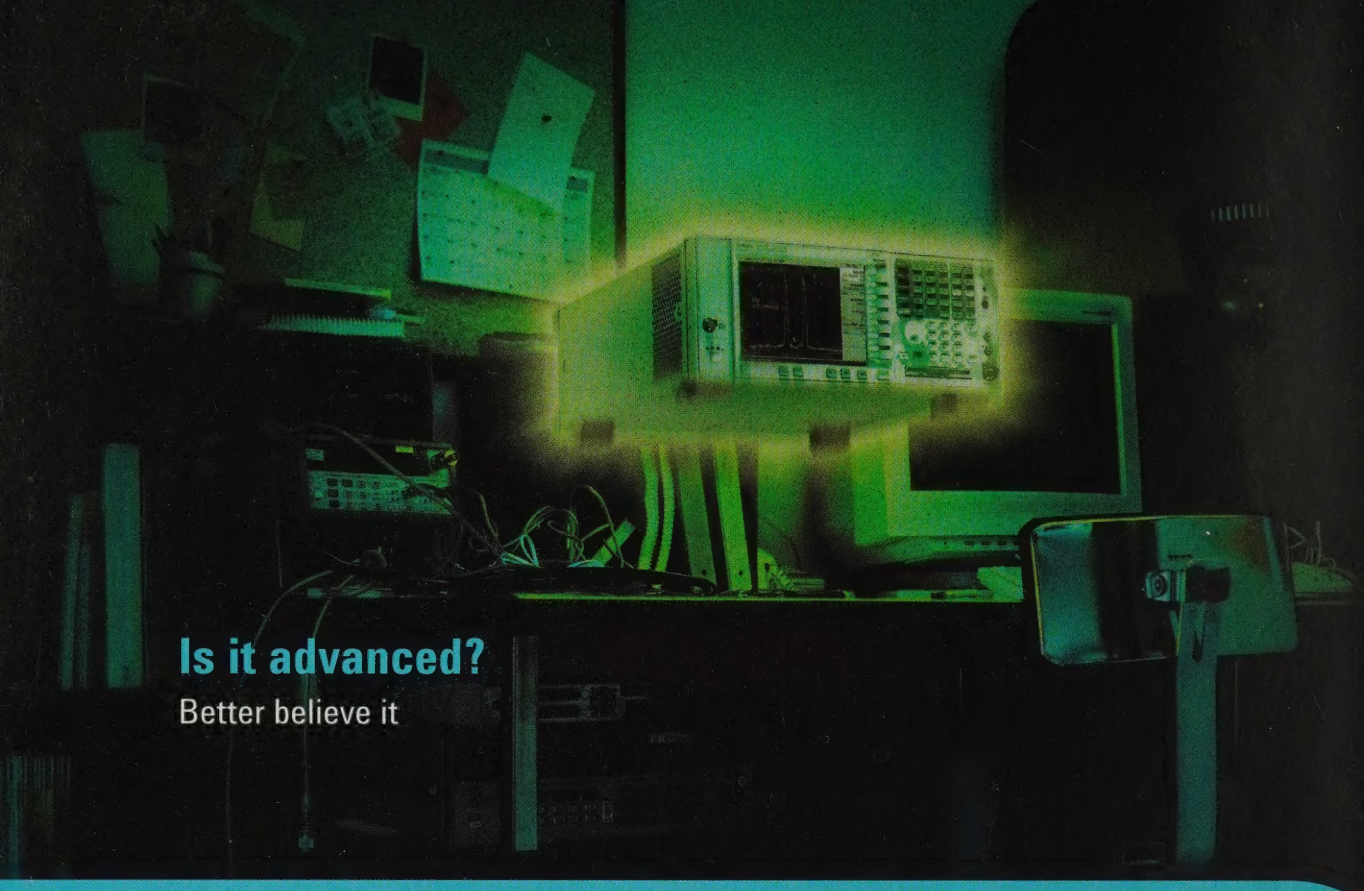
News
Tracing the history of an industry

Design Feature
Optimize designs for efficiency

Product Technology
3D process shrinks components/assemblies

Quad Modulators Stretch Dynamic Range To 4 GHz

Integrated Circuits Issue



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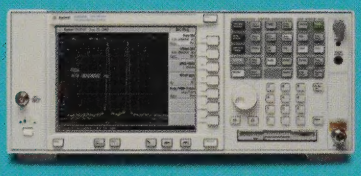

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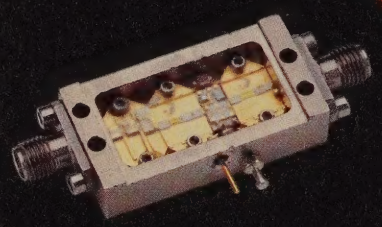

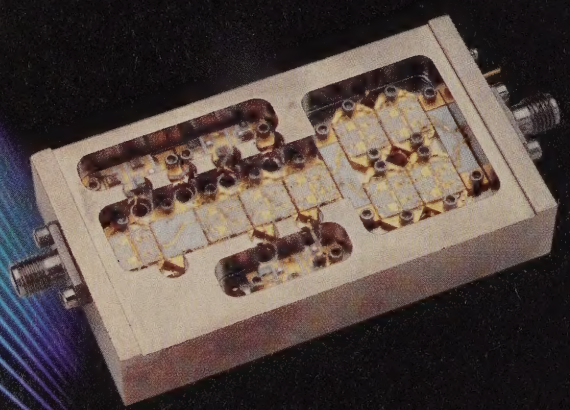

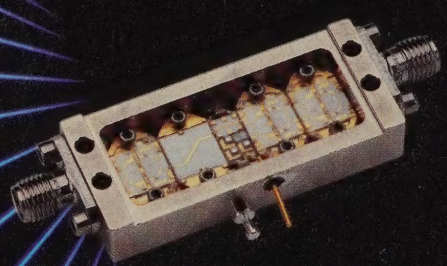

ULTRA BROAD BAND

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA018-203	0.5-18.0	20	5.0	2.5	7	17	2.0:1	250
JCA018-204	0.5-18.0	25	4.0	2.5	10	20	2.0:1	300
JCA218-506	2.0-18.0	35	5.0	2.5	15	25	2.0:1	400
JCA218-507	2.0-18.0	35	5.0	2.5	18	28	2.0:1	450
JCA218-407	2.0-18.0	30	5.0	2.5	21	31	2.0:1	500

MULTI OCTAVE AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA04-403	0.5-4.0	27	5.0	1.5	17	27	2.0:1	550
JCA08-417	0.5-8.0	32	4.5	1.5	17	27	2.0:1	550
JCA28-305	2.0-8.0	22	5.0	1.0	20	30	2.0:1	550
JCA212-603	2.0-12.0	32	5.0	3.0	14	24	2.0:1	550
JCA618-406	6.0-18.0	20	6.0	2.0	25	35	2.0:1	600
JCA618-507	6.0-18.0	25	6.0	2.0	27	37	2.0:1	800

MEDIUM POWER AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-P01	1.35-1.85	35	4.0	1.0	33	41	2.0:1	1000
JCA34-P02	3.1-3.5	40	4.5	1.0	37	45	2.0:1	2200
JCA56-P01	5.9-6.4	30	5.0	1.0	34	42	2.0:1	1200
JCA812-P03	8.0-12.0	40	5.0	1.5	33	40	2.0:1	1700
JCA1218-P02	12.0-18.0	22	4.0	2.0	25	35	2.0:1	700

LOW NOISE OCTAVE BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-3001	1.0-2.0	40	0.8	1.0	10	20	2.0:1	200
JCA24-3001	2.0-4.0	32	1.2	1.0	10	20	2.0:1	200
JCA48-3001	4.0-8.0	40	1.3	1.0	10	20	2.0:1	200
JCA812-3001	8.0-12.0	32	1.8	1.0	10	20	2.0:1	200
JCA1218-800	12.0-18.0	45	2.0	1.0	10	20	2.0:1	250

NARROW BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-1000	1.2-1.6	25	0.75	0.5	10	20	2.0:1	80
JCA23-302	2.2-2.3	30	0.8	0.5	10	20	2.0:1	80
JCA34-301	3.7-4.2	30	1.0	0.5	10	20	2.0:1	90
JCA56-401	5.4-5.9	40	1.0	0.5	10	20	2.0:1	120
JCA78-300	7.25-7.75	27	1.2	0.5	13	23	2.0:1	120
JCA910-3000	9.0-9.5	25	1.2	0.5	13	23	1.5:1	150
JCA910-3001	9.5-10.0	25	1.2	0.5	13	23	1.5:1	150
JCA1112-3000	11.7-12.2	27	1.1	0.5	13	23	1.5:1	150
JCA1213-3001	12.2-12.7	25	1.1	0.5	10	20	2.0:1	200
JCA1415-3001	14.4-15.4	35	1.4	1.0	14	24	2.0:1	200
JCA1819-3001	18.1-18.6	25	1.8	0.5	10	20	2.0:1	200
JCA2021-3001	20.2-21.2	25	2.0	0.5	10	20	2.0:1	200

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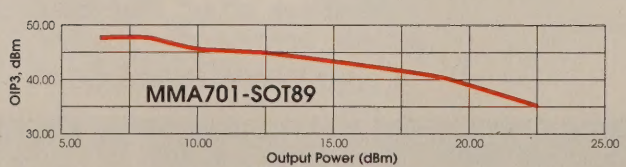
HBT AMPLIFIERS

Model	Frequency Range (GHz)	Vcc (V)	Icc (mA)	Gain* (dB)	P_{-1dB}** (dBm) Typical	IP3** (dBm) Typical	Thermal Resistance (°C/W)
MMA701-SOT89	0.001-4.0	7.0	130	12.5	+27	+48	80
MMA710-SOT89	0.001-5.0	7.0	95	12.5	+22	+37	130

* Matched SSG at 2.0 GHz
** Measured at 2.0 GHz



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This pair of direct quadrature modulators allows designers to eliminate a frequency-conversion stage in their BTS designs without sacrificing dynamic range.

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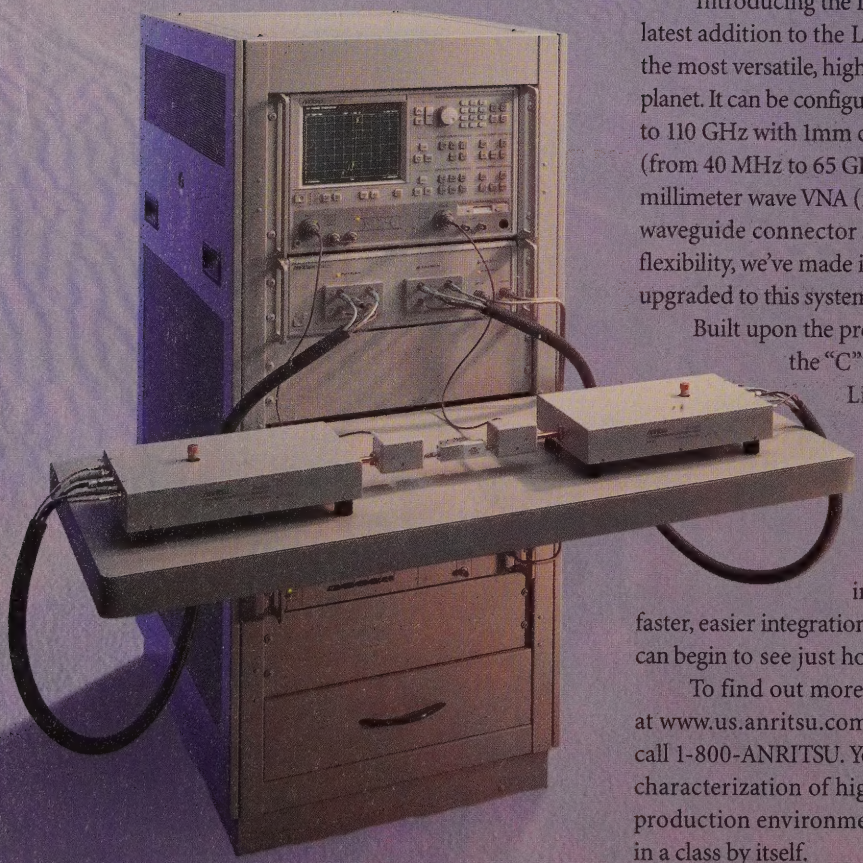

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ELECTRICAL SPECIFICATIONS

MODEL	FROM	TO	FREQ RANGE (GHz)	VSWR (GHz)
8006E1	QT3.5mm™ (m) with no nut	3.5mm (f)	DC — 26.5**	DC — 16.0, 1.05 16.0 — 26.5, 1.08
8006E11	QT3.5mm™ (m) with 3/8" dia. nut	3.5mm (f)		
8006E21	QT3.5mm™ (m) with 9/16" dia. nut	3.5mm (f)		
8006Q1	QT3.5mm™ (m) with guide sleeve	3.5mm (f)		

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REPEATABILITY	DC — 18.0 GHz	18.0 — 26.5 GHz
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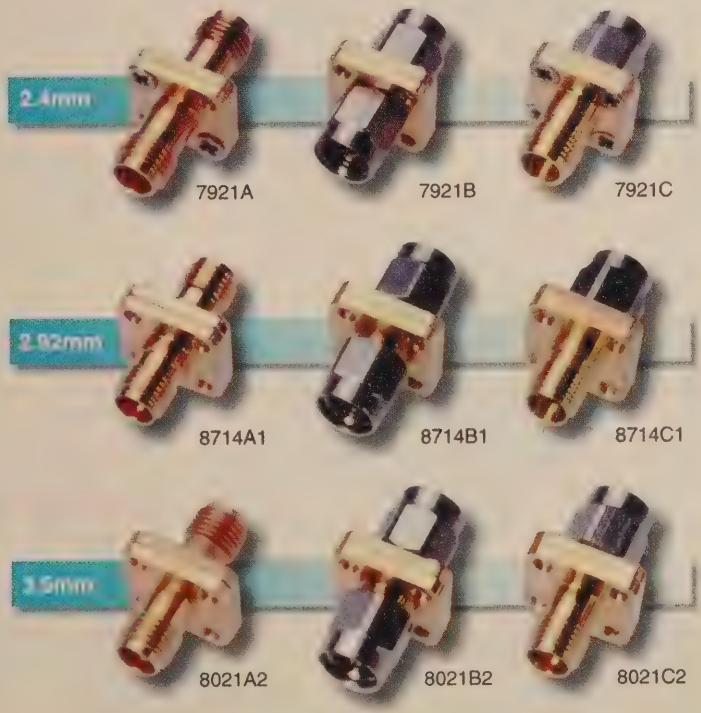

ELECTRICAL SPECIFICATIONS

MODEL	FROM	TO	FREQ RANGE & MAX. VSWR
7921A	2.4mm Q (f)	2.4mm Q (f)	DC — 26.5 GHz, 1.06 26.5 — 40.0 GHz, 1.10 26.5 — 34.0 GHz, 1.15
7921B	2.4mm Q (f)	2.4mm Q (m)	
7921C	2.4mm Q (f)	2.4mm Q (m)	
8714A1	2.92mm K (f)	2.92mm K (f)	DC — 4.0 GHz, 1.05 4.0 — 20.0 GHz, 1.08 20.0 — 40.0 GHz, 1.12
8714B1	2.92mm K (m)	2.92mm K (m)	
8714C1	2.92mm K (f)	2.92mm K (m)	
8021A2	3.5mm (f)	3.5mm (f)	DC — 18.0 GHz, 1.05 18.0 — 26.5 GHz, 1.08 26.5 — 34.0 GHz, 1.12
8021B2	3.5mm (m)	3.5mm (m)	
8021C2	3.5mm (f)	3.5mm (m)	

Between-Series configurations include: • 2.4mm to 2.92mm (K) • 2.4mm to 3.5mm

* U.S. PATENT #6210221

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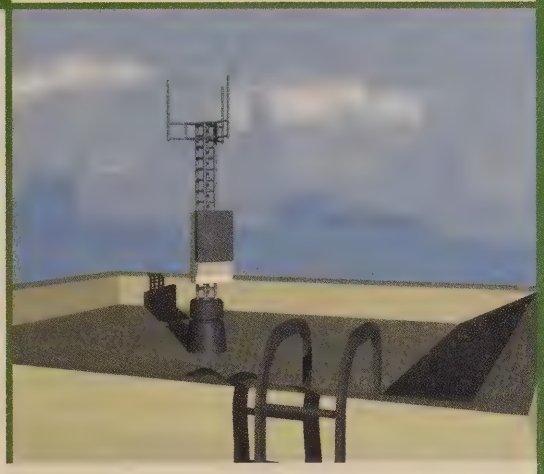

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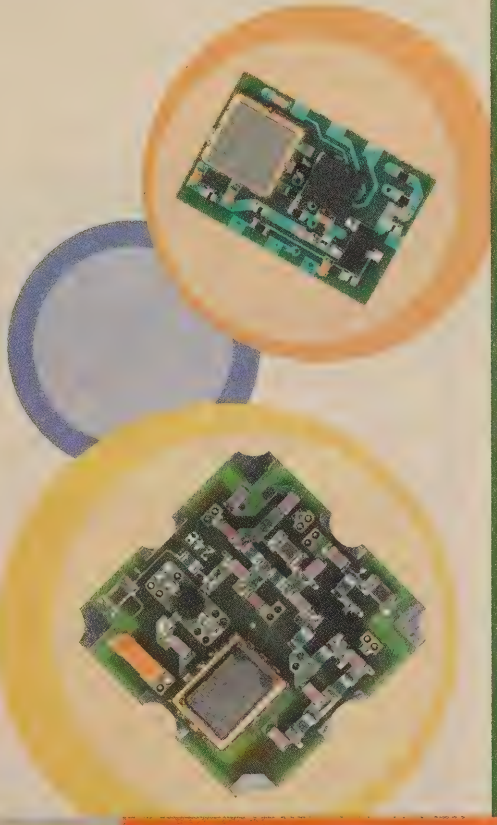

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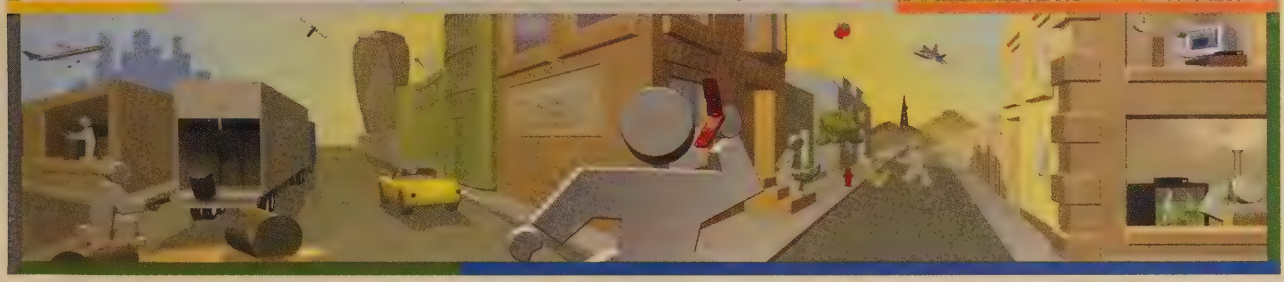

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Author Response

(continued from September's Feedback)

▶▶THE MEANING AND importance of the biphase line code is understood by noting that the signal is no longer a carrier with two islands, but becomes an SSB alone. It has been generated without a carrier ever having been present. This SSB can be transmitted and detected by the normal methods by using a phantom carrier. The fact that the signal appears to be a single frequency line would appear to violate Carson's rule. This is not the case, however, since inserting the phantom carrier restores the signal to the full bandwidth that is required by Carson's rule. It should be noted that since only a very-narrow bandwidth has been transmitted, the noise bandwidth at the Rx is less than that called for by Carson's rule and the Nyquist bandwidth. This is of great importance in the mathematical analysis of the signal.

In previous work of H.R. Walker, the encoder uses rectangular pulses (which is our belief) to spread the spectrum around the bit-rate frequency. While in our proposed algorithm, the encoder uses a sine wave that has a spike spectrum if zero intersection points are equal (this refers to a pure sine wave of fundamental frequency) and a little spike amplitude if they differ slightly. The benefit of this modulation format is its narrow spectrum and its simplicity in implementation, compared to the previous work of H.R. Walker.

It is well-known from basic signal theory that if zero intersection points of a sine wave are equal, then it is a pure sine wave of fundamental frequency. This means that the filter used in the Tx will need to be that of a bandpass type with zero bandwidth to pass the fundamental bit-rate frequency. This gives an indication that a significant reduction in the bandwidth of the occupied spectrum can be obtained.

The new algorithm of a VMSK system, as provided in our paper, was simulated and its exact theoretical results were given in Figs. 4 and 5. I believe that Mr. Karn has never played with the narrowband subject, using incomplete data.

K.H. Sayhood
Southeast University
Nanjing, People's Republic of China

Headline Error

▶▶IN THE AUGUST 2001 issue, the headline in one of your Product Technology stories is incorrect. On page 199, the headline reads "SAW Filter Screens PS Receive Signals." It should read "SAW Filter Screens GPS Receive Signals." The "G" in "GPS" was apparently dropped off.

Stephanie Orr
SAWTEK, Inc.

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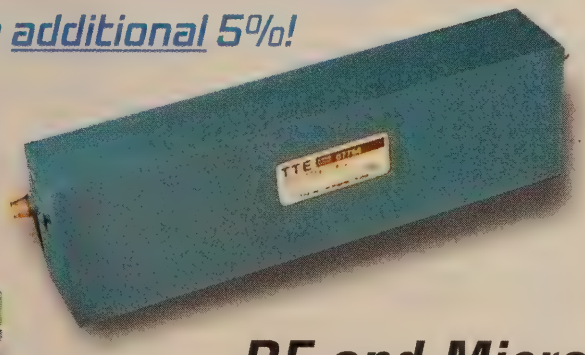

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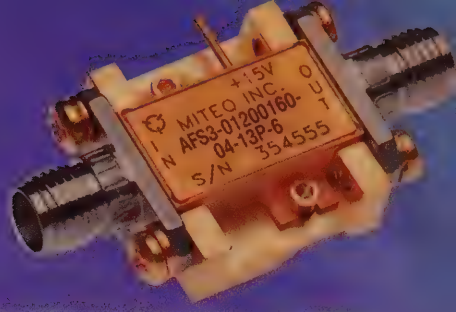

Model Number	Frequency Range (GHz)	Gain (Min./Max.) (dB)	Gain Flatness (±dB, Max.)	Noise Figure (dB, Max.)	VSWR Input (Max.)	VSWR Output (Max.)	Output Power @ 1 dB Comp. (dBm, Min.)	Nom. DC Power (+15 V, mA)
TEMPERATURE COMPENSATED AMPLIFIERS								
AFS3-01000200-15-TC-6	1–2	36-40	1.00	1.5	2.0:1	2.0:1	+5	125
AFS2-02000400-15-TC-6	2–4	22-26	1.00	1.5	2.0:1	2.0:1	+5	125
AFS3-02000400-15-TC-6	2–4	22-26	1.00	1.5	2.0:1	2.0:1	+5	125
AFS2-04000800-20-TC-2	4–8	18-22	1.00	2.0	2.0:1	2.0:1	+5	100
AFS3-04000800-18-TC-4	4–8	26-30	1.00	1.8	2.0:1	2.0:1	+8	150
AFS2-02000800-40-TC-2	2–8	14-19	1.50	4.0	2.0:1	2.0:1	+5	100
AFS3-02000800-30-TC-4	2–8	22-27	1.50	3.0	2.0:1	2.2:1	+8	150
AFS2-08001200-30-TC-2	8–12	12-16	1.00	3.0	2.0:1	2.0:1	+5	100
AFS3-08001200-22-TC-4	8–12	24-28	1.00	2.2	2.0:1	2.0:1	+8	150
AFS4-12001800-30-TC-8	12–18	22-26	1.00	3.0	2.0:1	2.0:1	+8	250
AFS4-06001800-35-TC-6	6–18	22-26	1.00	3.5	2.0:1	2.0:1	+8	250
AFS6-06001800-35-TC-8	6–18	30-34	1.00	3.5	2.0:1	2.0:1	+8	400
AFS4-02001800-45-TC-5	2–18	18-24	1.50	4.5	2.2:1	2.2:1	+8	175

Note: All specifications guaranteed -54 to +85°C.
Many other frequencies, noise figures and gain windows are available.

Model Number	Frequency Range (GHz)	Gain (Min./Max.) (dB)	Gain Flatness (±dB, Max.)	Noise Figure (dB, Max.)	VSWR Input (Max.)	VSWR Output (Max.)	Output Power @ 1 dB Comp. (dBm, Min.)	Nom. DC Power (+15 V, mA)
HIGHER POWER AMPLIFIERS								
AFS4-00050100-25-25P-6	0.5–2	36	1.50	2.5*	2.0:1	2.5:1	+25	325
AFS3-00100100-23-25P-6	.1–1	38	2.00	2.3	2.5:1	2.5:1	+25	280
AFS3-00100200-25-27P-6	.1–2	33	1.50	2.5	2.0:1	2.5:1	+27	300
AFS3-00100300-25-23P-6	.1–3	25	1.50	2.5	2.0:1	2.5:1	+23	300
AFS3-00100400-26-20P-4	.1–4	26	1.50	2.6	2.0:1	2.0:1	+20	250
AFS4-00100600-25-20P-4	.1–6	32	1.50	2.5	2.0:1	2.0:1	+20	300
AFS4-00100800-28-20P-4	.1–8	30	1.50	2.8	2.0:1	2.0:1	+20	300
AFS4-00101200-40-20P-4	.1–12	20	1.50	4.0	2.0:1	2.0:1	+20	300
AFS4-00501800-60-20P-6	.5–18	25	2.75	6.0	2.5:1	2.5:1	+20	350
AFS5-00102000-60-18P-6	.1–20	25	3.00	6.0	2.5:1	2.5:1	+18	360
AFS3-01000200-20-27P-6	1–2	33	1.50	2.0	2.0:1	2.0:1	+27	350
AFS3-02000400-30-25P-6	2–4	28	1.50	3.0	2.0:1	2.0:1	+25	250
AFS3-04000800-40-20P-4	4–8	20	1.00	4.0	2.0:1	2.0:1	+20	200
AFS4-08001200-50-20P-4	8–12	22	1.25	5.0	2.0:1	2.0:1	+20	200
AFS6-12001800-40-20P-6	12–18	28	2.00	4.0	2.0:1	2.0:1	+20	375
AFS6-06001800-50-20P-6	6–18	23	2.00	5.0	2.0:1	2.0:1	+20	365
AFS4-02001800-60-20P-6	2–18	23	2.50	6.0	2.5:1	2.0:1	+20	350

*Noise figure degrades below 100 MHz. Please consult factory for details.
Note: Noise figures increase below 500 MHz in bands wider than .1-10 GHz.

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Model Number	Frequency Range (GHz)	Gain (Min.) (dB)	Gain Flatness (±dB)	Noise Figure (dB, Max.)	VSWR Input (Max.)	VSWR Output (Max.)	Output Power @ 1 dB Comp. (dBm, Min.)	Nom. DC Power (+15 V, mA)
MODERATE BAND AMPLIFIERS								
AFS2-00700080-05-10P-4	.7–.8	30	0.50	0.45	1.5:1	1.5:1	+10	90
AFS2-00800100-05-10P-4	.8–1	30	0.50	0.45	1.5:1	1.5:1	+10	90
AFS3-01200160-05-13P-6	1.2–1.6	40	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-01400170-05-13P-6	1.4–1.7	40	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-01500180-04-13P-6	1.5–1.8	40	0.50	0.40	1.5:1	1.5:1	+13	150
AFS3-01500250-06-13P-6	1.5–2.5	36	0.50	0.60	2.0:1	2.0:1	+13	150
AFS3-01700190-04-13P-6	1.7–1.9	36	0.50	0.40	1.5:1	1.5:1	+13	150
AFS3-01800220-05-13P-6	1.8–2.2	36	0.50	0.50	1.5:1	1.5:1	+13	150
AFS3-02200230-04-13P-6	2.2–2.3	36	0.50	0.40	1.5:1	1.5:1	+13	150
AFS3-02300270-05-13P-6	2.3–2.7	34	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-02700290-05-13P-6	2.7–2.9	32	0.50	0.50	1.5:1	1.5:1	+13	150
AFS3-02900310-05-13P-6	2.9–3.1	32	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-03100350-06-10P-4	3.1–3.5	29	0.50	0.6	1.5:1	1.5:1	+10	150
AFS4-03400420-06-13P-6	3.4–4.2	40	0.50	0.60	1.5:1	1.5:1	+13	225
AFS3-04400510-07-5P-4	4.4–5.1	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-04500480-07-5P-4	4.5–4.8	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-05200600-07-5P-4	5.2–6	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-05400590-07-5P-4	5.4–5.9	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-05800670-07-5P-4	5.8–6.7	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-07250775-06-5P-4	7.25–7.75	30	0.50	0.60	1.5:1	1.5:1	+5	100
AFS3-07900840-07-5P-4	7.9–8.4	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS4-08500960-08-5P-4	8.5–9.6	32	0.75	0.80	1.5:1	1.5:1	+5	125
AFS3-09001100-09-5P-4	9–11	26	0.50	0.90	1.5:1	1.5:1	+5	100
AFS4-09001100-09-5P-4	9–11	32	0.75	0.90	1.5:1	1.5:1	+5	125
AFS4-10951175-09-5P-4	10.95–11.75	32	0.75	0.90	1.5:1	1.5:1	+5	125
AFS4-11701220-09-5P-4	11.7–12.2	32	0.75	0.90	1.5:1	1.5:1	+5	125
AFS2-12201280-10-8P-4	12.2–12.8	14	0.75	1.00	1.5:1	1.5:1	+8	80
AFS4-12201280-10-12P-4	12.2–12.8	27	0.75	1.00	1.5:1	1.5:1	+12	200
AFS4-12701330-13-10P-4	12.7–13.3	27	0.75	1.30	1.5:1	1.5:1	+10	175
AFS4-13201400-14-10P-4	13.2–14	24	0.75	1.40	1.5:1	1.5:1	+10	175
AFS4-14001450-14-10P-4	14–14.5	24	0.75	1.40	1.5:1	1.5:1	+10	175
AFS4-20202120-20-8P-4	20.2–21.2	20	1.00	2.00	1.5:1	1.5:1	+8	175
AFS4-21202400-22-10P-4	21.2–24	18	1.00	2.2	2.0:1	2.0:1	+10	100
OCTAVE BAND AMPLIFIERS								
AFS3-00120025-09-10P-4	.12–.25	38	0.50	0.9	2.0:1	2.0:1	+10	175
AFS3-00250050-08-10P-4	.25–.5	38	0.50	0.8	2.0:1	2.0:1	+10	125
AFS3-00500100-05-10P-6	.5–1	38	0.75	0.5	2.0:1	2.0:1	+10	150
AFS3-01000200-05-10P-6	1–2	38	1.00	0.5	2.0:1	2.0:1	+10	150
AFS3-01200240-05-10P-6	1.2–2.4	34	1.00	0.5	2.0:1	2.0:1	+10	175
AFS3-02000400-06-10P-4	2–4	30	1.00	0.6	2.0:1	2.0:1	+10	125
AFS3-02600520-10-10P-4	2.6–5.2	28	1.00	1.0	2.0:1	2.0:1	+10	150
AFS3-04000800-07-10P-4	4–8	30	1.00	0.7	2.0:1	2.0:1	+10	125
AFS3-08001200-09-10P-4	8–12	26	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-08001600-15-8P-4	8–16	26	1.00	1.5	2.0:1	2.0:1	+8	80
AFS4-12002400-25-10P-4	12–24	20	2.00	2.5	2.0:1	2.0:1	+10	85
AFS4-12001800-18-10P-4	12–18	26	1.00	1.8	2.0:1	2.0:1	+10	125
AFS4-18002650-28-8P-4	18–26.5	18	1.75	2.8	2.5:1	2.2:1	+8	150
MULTIOCTAVE BAND AMPLIFIERS								
AFS1-00040200-12-10P-4	.04–2	15	1.50	1.2	2.5:1	2.0:1	+10	75
AFS3-00300140-08-10P-4	.3–1.4	33	1.00	0.8	2.0:1	2.0:1	+10	150
AFS2-00400350-12-10P-4	.4–3.5	22	1.50	1.2	2.0:1	2.0:1	+10	80
AFS3-00500200-08-15P-4	.5–2	38	1.00	0.8	2.0:1	2.0:1	+15	125
AFS3-01000400-09-10P-4	1–4	30	1.50	0.9	2.0:1	2.0:1	+10	125
AFS3-02000800-09-10P-4	2–8	26	1.00	0.9	2.0:1	2.0:1	+10	125
AFS4-02001800-23-10P-4	2–18	25	2.00	2.3	2.0:1	2.0:1	+10	175
AFS4-06001800-22-10P-4	6–18	24	2.00	2.2	2.0:1	2.0:1	+10	150
AFS4-08001800-22-10P-4	8–18	26	2.00	2.2	2.0:1	2.0:1	+10	150
ULTRA WIDEBAND AMPLIFIERS								
AFS3-00100100-09-10P-4	.1–1	38	1.00	0.9	2.0:1	2.0:1	+10	150
AFS3-00100200-10-15P-4	.1–2	38	1.00	1.0	2.0:1	2.0:1	+15	150
AFS3-00100300-11-10P-4	.1–3	32	1.00	1.1	2.0:1	2.0:1	+10	150
AFS3-00100400-13-10P-4	.1–4	28	1.00	1.3	2.0:1	2.0:1	+10	150
AFS3-00100600-13-10P-4	.1–6	28	1.25	1.3	2.0:1	2.0:1	+10	125
AFS3-00100800-14-10P-4	.1–8	25	1.50	1.4	2.0:1	2.0:1	+10	125
AFS4-00101200-22-10P-4	.1–12	28	1.50	2.2	2.0:1	2.0:1	+10	175
AFS4-00101400-23-10P-4	.1–14	24	2.00	2.3	2.5:1	2.5:1	+10	200
AFS4-00101800-25-10P-4	.1–18	25	2.00	2.5	2.5:1	2.5:1	+10	175
AFS4-00102000-30-10P-4	.1–20	20	2.50	3.0	2.5:1	2.5:1	+10	175
AFS4-00102650-40-8P-4	.1–26.5	18	2.50	4.0	2.5:1	2.5:1	+8	175

Note: Noise figure increases below 500 MHz in bands greater than 0.1-10 GHz.

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The Face Of War Has Changed

HORROR MIGHT BE the single best word to describe what most of the civilized world felt this past month, when hijacked commercial aircraft were flown into Washington DC's Pentagon building and the twin towers of New York City's World Trade Center. For those who lost loved ones in those aircraft, the rubble of the buildings, the airliner that crashed in Pennsylvania, or in the rescue efforts, the tragedy will last a lifetime. And for those of us standing by, watching helplessly, there is a sense of deep sorrow in knowing the world had changed forever that morning of September 11, 2001.

It is up to this nation and this industry to apply ingenuity, to invest in our Armed Forces, and help them make the transition to a more mobile fighting machine.

For those of us in engineering, there is a tendency to analyze what happened that morning. How could it have been so easy to commandeer four planes and use them as weapons against the American people? We live in an age of sophisticated electronics, capable of measuring the faintest signals. Was our intelligence so lacking that these acts of terrorism or, more precisely, these acts of war were so unexpected? Several days after these tragic events, transcripts from the trial for the terrorists of the first attempt to destroy the World Trade Center, some eight years ago, revealed statements by the accused that it would be a simple matter to rent an aircraft and blow up the CIA.

America is the most technologically sophisticated country in the world. But we have spent the last decade ignoring our military and paying too much attention to stock-market prices. The attack on democracy this past September 11th was just a hint of things to come, a wake-up call for those willing to be roused. We have been at war for some time, with a cunning and ruthless enemy. And we have underestimated that enemy too many times, only because they are based in such primitive places as Afghanistan and Iraq.

It is up to this nation and this industry to apply their ingenuity, to invest in our Armed Forces, and to help them make the transition to a more mobile, more "terrorist-ready" fighting machine. The enemy will not go away—their hatred of America and democracy is too great. Due to the fact that the face of war has changed greatly from what we have known, we must unite as one to win that war.

If we need a reminder, let the memory of the World Trade Center crumbling to the ground remain strong in our minds. Let the anguish of those who lost friends and family in those buildings and aircraft remind us, and let the tremendous bravery of the rescue teams—firefighters, police officers, government workers, and volunteers—serve as the inspiration for what this nation can accomplish when it stands together.

Jack Browne

Publisher/Editor

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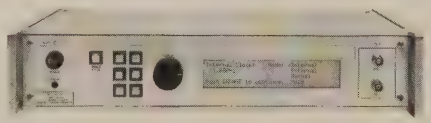

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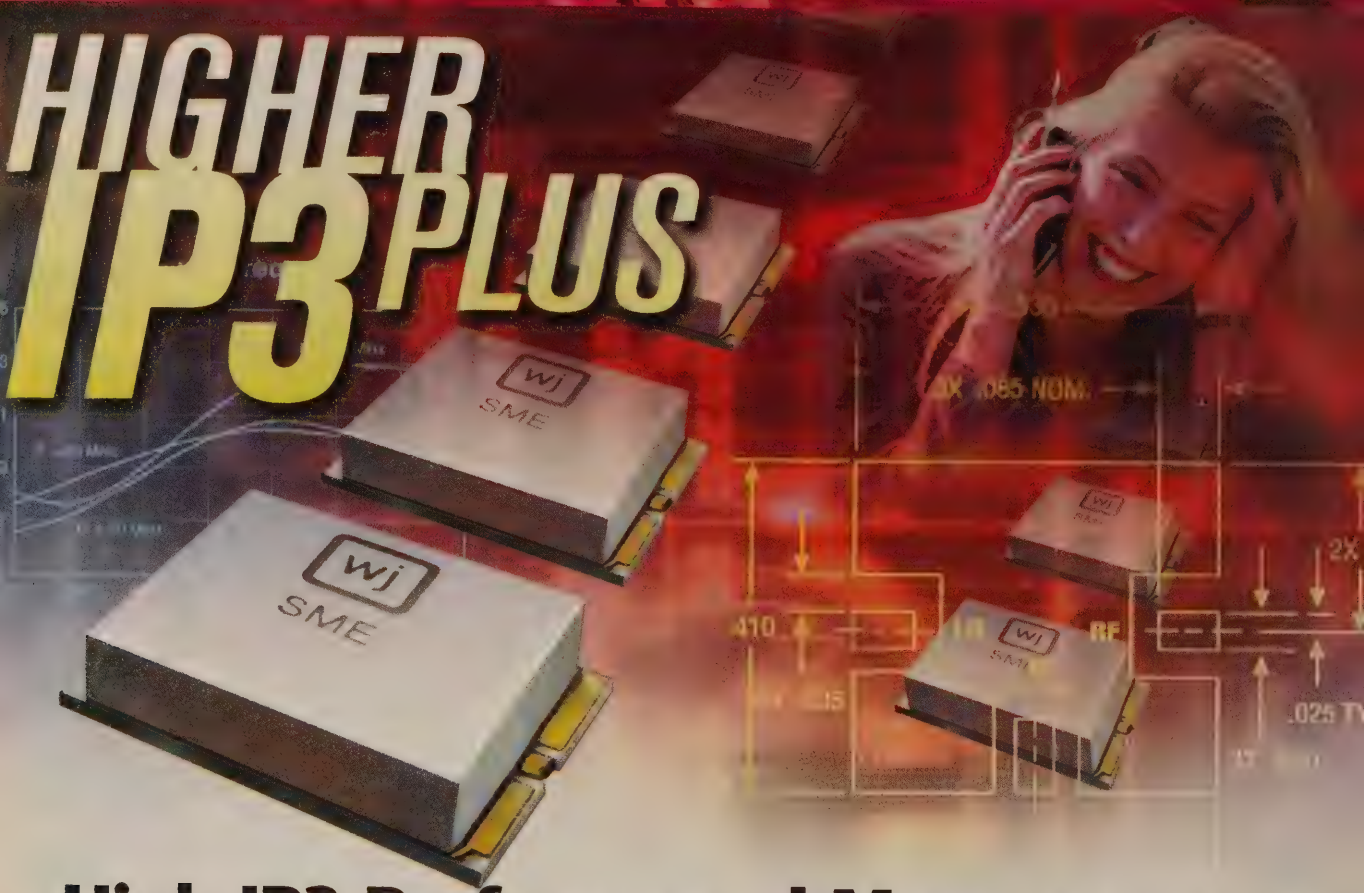

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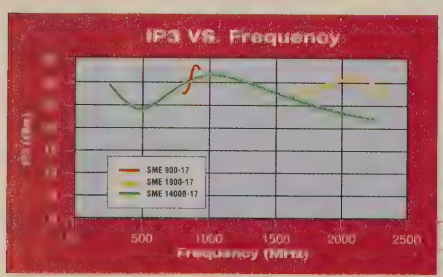

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SME 1400B-10	1-2200	1-2200	1-2000	+10	+6	+19	6.5	30
SME 1400B-13	1-2200	1-2200	1-2000	+13	+9	+22	6.5	30
SME 1400B-17	1-2200	1-2200	1-2000	+17	+13	+27	6.5	30
SME 1900-17	1600-2400	1400-2390	10-250	+17	+14	+29	7.4	26

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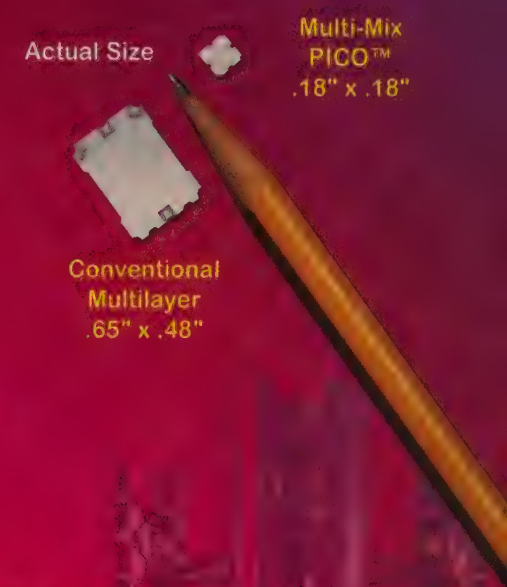

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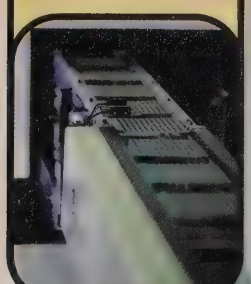

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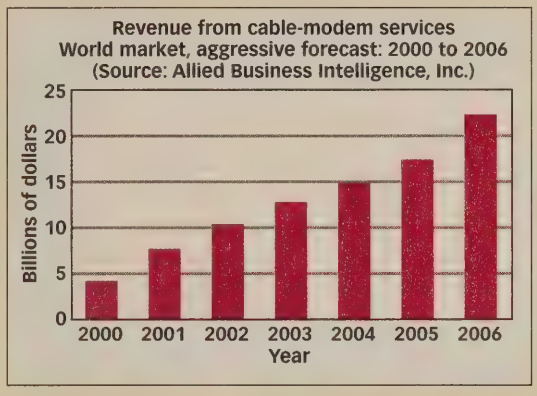
Cable Still Holds Advantage Over DSL

OYSTER BAY, NY—Due to its head start in technological development, cable-based broadband services will prevail over telephone-wire-based digital subscriber line (DSL) in the battle for dominance among wire-line broadband technologies, according to an Allied Business Intelligence (ABI) study entitled, "Cable vs. DSL: The Race Is On."

Revenues from cable broadband-access subscriber fees will triple in the next five years, from 7.5 billion this year to $22.2 billion in 2006 **(see figure)**. The subscriber base will grow from 14.2 million to 48.7 million for the same period.



"DSL providers have been sitting on their technology for too long, and in the last three years, have been in too much of a race to catch up with cable access to satisfactorily resolve issues of supply, deployment, and shared infrastructure access," says Mark M. Fox, ABI broadband analyst and the report's author. "They will certainly regroup, and DSL will still be a major player in the wire-line broadband market, but by then, cable operators will have a firm grip on the market."

Cable-modem technology will capitalize on its technological edge over the copper (Cu)-wire-based DSL. Not only can hybrid fiber coaxial cable, predominant in modern networks, carry hundreds of times more data than a twisted pair of telephone wires, but there are relatively inexpensive solutions to eliminate the problem of interference that is created by the shared cable network. One solution is to split the existing node so that each new node would serve fewer households. Another solution is to install new frequency-modulation (FM) equipment.

InP Foundry Services Have Become Available

CAMARILLO, CA—Vitesse Semiconductor Corp. has announced the availability of its indium-phosphide (InP) foundry services for designers of next-generation high-performance integrated circuits (ICs) and optoelectronic ICs (OEICs). Vitesse's indium phosphide (VIP) is suitable for fiber-based networks, including Synchronous Optical Network (SONET) OC-768 and 100-Gb/s Ethernet, wireless and RF applications, or any application that requires the mutual benefits of very-high-gain bandwidth product and high breakdown voltage.

Vitesse's proprietary InP process is based on a vertical NPN bipolar transistor. A single-heterojunction-bipolar-transistor (SHBT) process is in preproduction now with a double-HBT (DHBT) process, with higher breakdown voltage, to be made available soon. A third variation that is suitable for the integration of detectors with electronics is also available. An F_t of 150 GHz and an F_{max} of 150 GHz at a breakdown voltage of +4.2 VDC is available in the SHBT process. Three layers of aluminum (Al)-interconnect metal, precision resistors, and metal-insulator-metal (MIM) capacitors complete this process.

"Our InP process is ideally suited to the fabrication of ICs with high breakdown voltage and fast edge rates needed for 40-plus Gb/s network equipment. Peak transistor performance is achieved at low current density for superior thermal management," states Alan Huelsman, director of Vitesse's InP program.

MSM IC Is Integrated Into Samsung Handset

SAN DIEGO, CA—Qualcomm, a firm that is involved in code-division-multiple-access (CDMA) digital wireless technology; together with SchlumbergerSema, a smart-card technology company; and Samsung Telecommunications America, Inc., an electronics firm, have announced the successful demonstration of a removable user-identity-module (R-UIM) solution. The QCT-SchlumbergerSema-Samsung demonstration of the ability to roam across CDMA and Global System for Mobile Communications (GSM) networks was performed at the recent 3G World Congress in Hong Kong.

We have more than 100 Bluetooth platforms operating here in a confined space at the same time, and there have been no interference problems whatsoever."

The demonstration used a Qualcomm CDMA Technologies (QCT) Mobile Station Modem (MSM™) with an integrated R-UIM interface and R-UIM Simera Airflex™ system by SchlumbergerSema in a Samsung handset. This solution supports roaming services and smart-card capabilities across CDMA and GSM networks. The end-to-end system provides CDMA handset manufacturers with a comprehensive solution that allows a subscriber's identity to be stored in a single removable card, a capability that is mandatory for the Chinese wireless market. This system solution will make global roaming easier for countries using different frequencies, or across CDMA or GSM networks, by allowing subscribers to exchange handsets while using the same card to maintain all of their personal information, including network settings, privileges, preferences, phone books, and more.

"The incorporation of OCT's integrated circuit and system software with the SchlumbergerSema Simera Airflex card in Samsung's CDMA handsets for the Chinese market clearly demonstrates how industry-leading system providers can collaborate to rapidly deliver solutions for a key market," says Luis Pineda, vice president of product management for Qualcomm CDMA Technologies. "These solutions can now be adopted to roll out R-UIM-enabled handsets using our MSM products and SchlumbergerSema smart cards as CDMA network operators begin to specify this feature for their wireless devices."

"Samsung is very proud to have demonstrated our R-UIM-enabled handset at the CDMA Development Group's 3G World Congress," says Kim HyunJoo, research-and-development (R&D) manager of mobile communications for Samsung. "The system solutions from QCT and SchlumbergerSema, designed to meet the market requirements in China and Korea, allowed us to achieve significant time-to-market advantages to roll out a validated R-UIM solution."

The Sixth Bluetooth UnPlugFest Is A Big Success

SOPHIA ANTIPOLIS, FRANCE—The sixth Bluetooth UnPlugFest, hosted by the European Telecommunications Standards Institute (ETSI) from August 19 to 24, was given high marks by participants. The event, which was held at the Acropolis Congress Centre in Nice, France, attracted 440 engineers from 89 companies in 20 different countries.

The Bluetooth UnPlugFests allow companies to test the interoperability of their upcoming Bluetooth products. Bluetooth wireless technology is a specification for a small form factor, low-cost radio technology that can be integrated into mobile devices.

The first Bluetooth products hit the market at the end of last year. Attendance at the Bluetooth UnPlugFests has been rising steadily, demonstrating the momentum that Bluetooth wireless technology has generated and the high regard in which ETSI's interoperability services are held throughout the world. Thirty percent of participants at the latest event came from North America, 20 percent from Asia (mostly Japan), and 48 percent from Europe. Other countries, such as Australia, were also represented.

Intel's Mike Wilson, chairman of the Bluetooth SIG's UnPlugFest planning committee, says, "Many of the companies which previously brought experimental designs to the earlier events for testing are now coming with almost complete products." Wilson believes that future events may well include an extra phase of Bluetooth lifecycle—extending the existing proving of prototype designs to testing the interoperability between fully commercial products from the different participating companies.

This maturing of the technology was also noted by Nokia's Thomas Will, technical chairman of the UnPlugFest Steering Committee of the SIG. "We have more than 100 Bluetooth platforms operating here in a confined space at the same time, and there have been no interference problems whatsoever," notes Will.

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TST0950	900-MHz LNA	GSM, ISM
TST0912	900-MHz PA	GSM
TST0951	1900-MHz SiGe LNA	DCS & PCS mobile phones
T7024	2.4-GHz SiGe Front End	ISM/Bluetooth
T0980	400/500-MHz SiGe Front End	Family radio (Walky Talky) & remote control applications

PA: Power Amplifier
LNA: Low Noise Amplifier

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NIST Gives Chip Makers The Edge By Smoothing Edges

GAITHERSBURG, MD—Imagine trying to measure something very tiny that is also not very straight. That is the challenge faced by the semiconductor industry as computer chips become smaller. Miniaturization techniques have advanced considerably in recent decades. Yet it is difficult to come up with a good "ruler" for measuring something that is only 100 nm across, or approximately one one-thousandth the width of a human hair. The task is tougher yet if one is trying to measure something with ragged edges.

According to industry analysts, shipments of wireless handsets will exceed 700 million units by 2005."

Engineers at the National Institute of Standards and Technology (NIST) are helping the semiconductor industry overcome this problem and clear the way to improved measurements of "critical dimensions" (CD). CD refers to the smallest size that can be etched into a computer chip uniformly. Ragged-edged chip materials cannot provide a reliable, consistent standard for measuring CD. Therefore, NIST engineers have developed test structures using a silicon (Si) crystal lattice that has perfectly even sides to do the job.

With the introduction of these test structures, NIST has taken a step toward creating a standard way to calibrate instruments (the "rulers") used to measure 100-nm-wide gates in computer chips. NIST distributed test chips—each containing a prototype calibration standard with a CD ranging from 80 to 140 nm—to 15 semiconductor manufacturers (members of the International SEMATECH consortium) worldwide in April. The firms have evaluated the samples and provided the results to NIST last month. The data are now being used to complete work on a CD measurement standard for the semiconductor industry.

Velocium Opens 50,000-Sq.-Ft. Headquarters

EL SEGUNDO, CA—Velocium, TRW's telecommunication-products company, has opened a new headquarters building and product-design center. The 50,000-sq.-ft. facility is only minutes away from TRW's recently completed production line for fabricating high-performance indium-phosphide (InP) chips for telecommunications applications.

"The new headquarters demonstrates the growing emergence of Velocium as a TRW brand," says Dwight Streit, Velocium's president. "We're bringing TRW's advanced technology to market just when the telecommunications industry is seeking the high-speed optoelectronic solutions possible with indium phosphide."

Velocium is currently working jointly with Hitachi to develop InP power-amplifier (PA) modules for third-generation (3G) wireless communications.

According to industry analysts, shipments of wireless handsets will exceed 700 million units by 2005. 3G wireless units, together with 2.5G phones, are estimated to capture more than 50 percent of these shipments. All of the units will contain one or more PA chips.

Kudos

Ansoft Corp. announced that it has been added to the Russell® 2000 Index. The Frank Russell Co. adjusts its indexes annually in order to accurately rank the 3000 largest companies in the US stock market by market capitalization, known as the Russell 3000®. The Russell 2000 Index, representing the 2000 smallest stocks in the Russell 3000, is a leading benchmark for the small cap market...LeCroy Corp. has received a Quality Supplier Award from Cardinal Components...M2 Global, Inc., a worldwide supplier of microwave components and value-added services for the telecommunications and medical-equipment markets, has announced that its chairman of the board, Douglas Carlberg, has been reappointed to the Shingo Prize Board of Governors. The Shingo Prize was established in 1988 to promote awareness of lean manufacturing concepts and recognize companies that achieve world-class manufacturing status, customer satisfaction, and business results...A surprise gift of 47 dish antennas, feed assemblies, and antenna mounts has now brought The SETI League one step closer to realizing its long-planned phase-array telescope design. The donation, from father-and-son computer dealers Lionel and Jason Lumaghini of Rockland, MA and their respective companies (JCM Computers, LLC and Micro Technology), provides SETI League engineers with enough commercial-grade antennas and support hardware to test design concepts, and to eventually assemble their *Array2k* antenna project that was announced in 1999. MRF

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NGA-286	DC–6.0	15	31.2	15	3.4	4.0, 50	120
NGA-386	DC–4.0	15	27	19	2.7	4.0, 35	144
NGA-486	DC–8.0	17.5	39.5	14.5	4.5	4.2, 80	118
NGA-586	DC–6.0	19	38	19	4.5	5.0, 80	121
NGA-686	DC–4.0	19.2	35	11	6.1	5.9, 80	121

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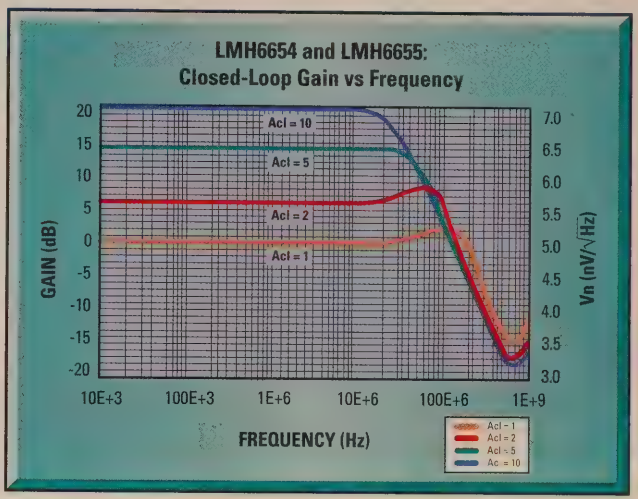

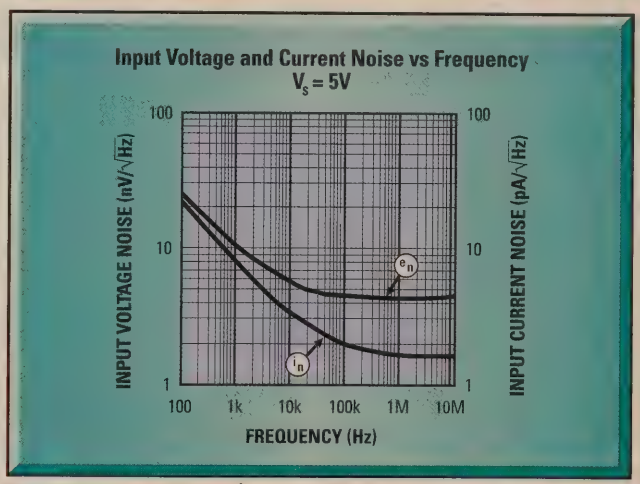

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NEWS

Tracing The History Of An Industry

The stories of these companies tell such things as the strength of a handshake, the flexibility to adapt to changing technologies, and how success can be learned.

History was made this past August when this magazine reached a milestone, celebrating its 40th year of publication. That issue also included the first in a series of histories about individual companies [see the article on ARRA, Inc. (Bay Shore, NY) in that issue, appearing on pp. 160-164]. But this industry is rich with the stories of many companies, and some of these stories will be told in the next several months, beginning with three of the longer-standing and more successful high-frequency firms: Alpha Industries (Woburn, MA), California Eastern Laboratories (Santa Clara, CA), and Rogers Corp. (Rogers, CT).

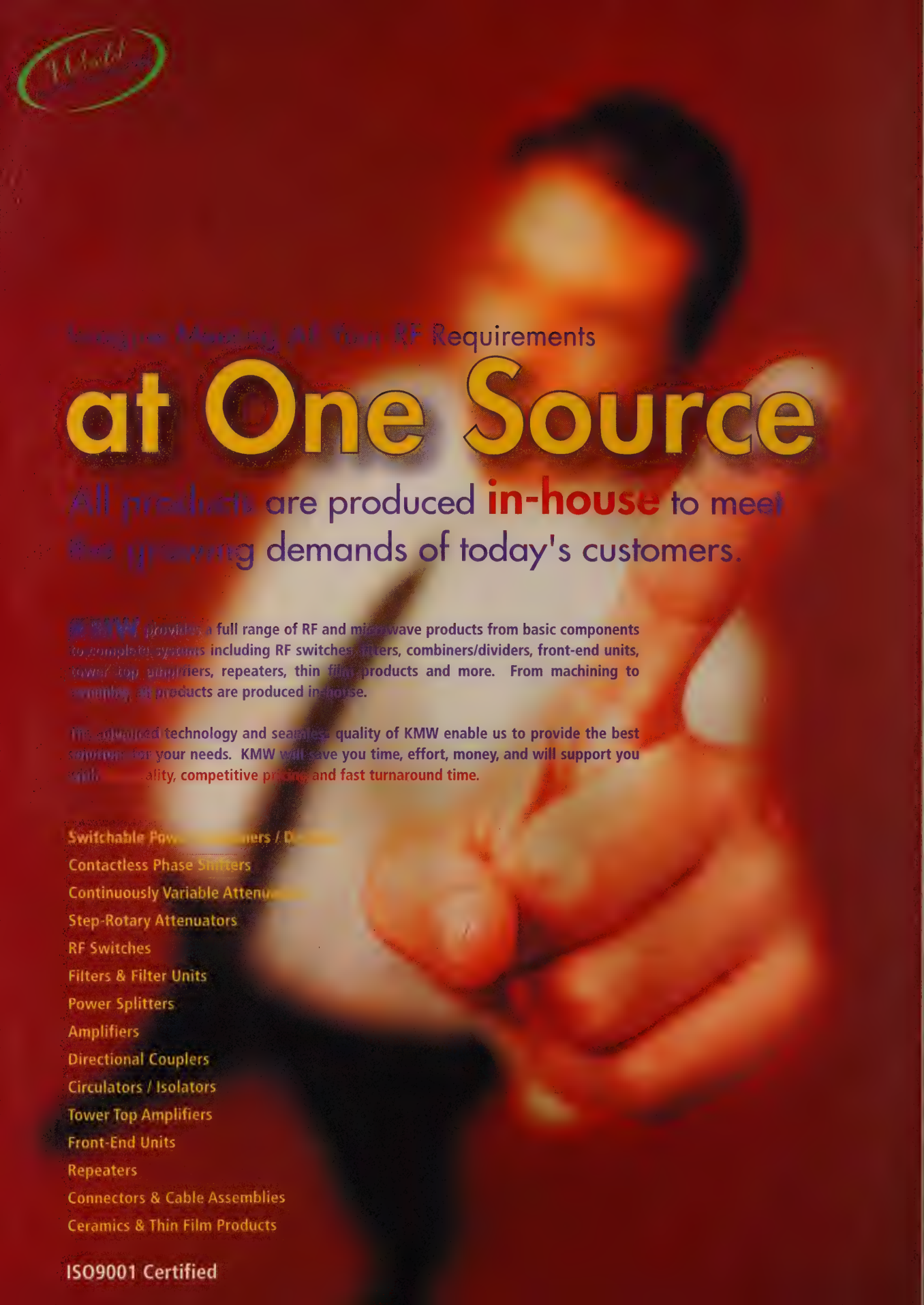
JACK BROWNE
Publisher/Editor

Alpha Industries

Alpha Industries began in an old textile mill with George Kariotis, his brother Andrew, and two other founding members in 1962 as Alpha Microwave. The founders worked for approximately one year without pay, building microwave waveguide assemblies for electronic-warfare (EW) systems. In 1963, the company began to fabricate microwave point-contact diodes. In 1965, following rapid growth, the name was changed to Alpha Industries, and two divisions were formed—the Microwave Components Division (with the original waveguide assemblies and a new diode-switch line) and the Semiconductor Division. Alpha began trading publicly in 1967.

One of the more significant events in Alpha's history was the acquisition of Sylvania's microwave-diode operation in 1970. The company added millimeter-wave capability in 1974 with the acquisition of TRG, and added an amplifier/thick-film operation known as Optimax that same year (which was eventually divested in 1987). Alpha expanded further in 1981 with the acquisitions of millimeter-wave-diode specialist Central Microwave Co. and ceramic-supplier Trans-Tech, Inc. Alpha's most recent acquisition, Network Device, Inc. (Sunnyvale, CA), brought Alpha foundry and design capabilities in gallium-arsenide (GaAs) heterojunction-bipolar-transistor (HBT) technology. Alpha is now one of the world's leading suppliers of microwave semiconductors, with more than 1000 employees in four facilities in Woburn, MA; Haverhill, MA; Adamstown, MD; and Sunnyvale, CA. Last year's revenues of $272 million marked a 46-percent growth in sales compared to 1999, and a 90-percent growth in operating income compared with the previous year's results.

Paul Vincent, Vice President and



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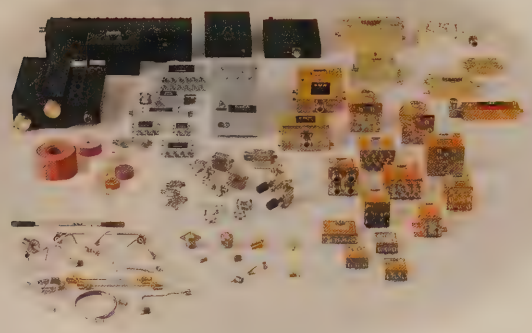

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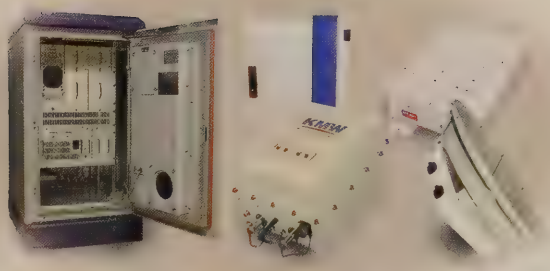

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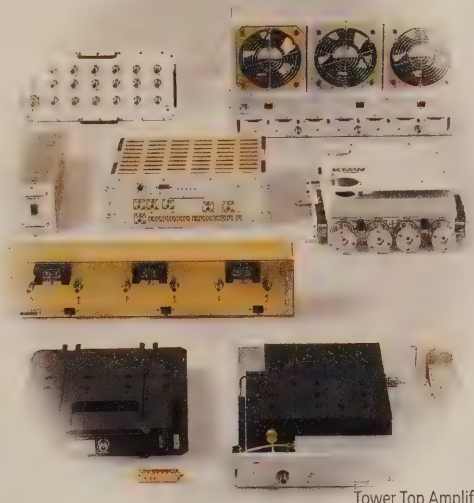

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CFO of Alpha Industries, joined Alpha in 1979 as Corporate Comptroller **(Fig. 1)**. He remained in that position until being named CFO in January 1997. One of the greatest challenges that faced the company during his tenure was the transition from a largely military base of customers to a commercial base. One key to making that transition was the major investment in semiconductor-processing equipment: "In the mid-1980s, under the leadership of Woody Reed, we invested about $20 million to start a GaAs fabrication facility, which was a considerable investment at that time for a company that was only $70 million in annual sales," noted Vincent. "Taking that risk allowed us to participate in the DARPA MIMIC program in the later part of the 1980s, in which we were partnered with Martin Marietta and ITT. Martin Marietta was the integrator. ITT brought the microwave technology and we brought the millimeter-wave technology.

"The shift from military to commercial was a fundamental shift not only from the products that we would produce but from the way that we would produce them," remembers Vincent. "When Tom Leonard [currently Chairman] joined Alpha in 1992 as General Manager of the Components and Subsystems Division, he got us to rethink how we were going to approach new markets with high-volume manufacturing. Tom helped us develop that whole manufacturing expertise for making high-volume GaAs products, so that when the large telecommunications markets finally arrived, we were in a good position to take advantage of those markets."

1. Paul Vincent, who joined Alpha Industries in 1979, is now Vice President and CFO.

According to Vincent, "Tom Leonard joined the company in 1991 or 1992, to run our Methuen, MA operation, which was consolidated with the Woburn operation in 1994. Tom's first action was to train everyone in his group, including me, about the TQM process. And this was not just a short course—this training took place over a eight-week period. Tom then expanded the training to all employees.

"We've always been good communicators," notes Vincent, "and the company has always been set up in a nonhierarchical way so that there would be strong communications between people at all levels in the company. And this goes back to George Kariotis, who used to exclaim that

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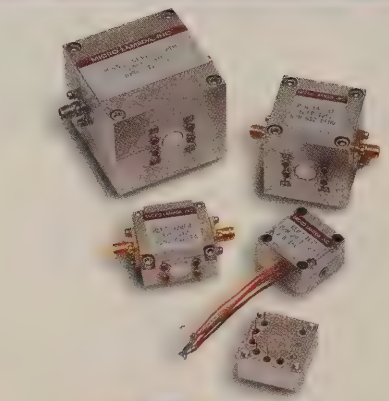

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2. Jerry Arden is President of California Eastern Labs.

'there would be no neckties in this company.' And that was in the 1960s when the culture was still very strong for wearing suits and ties in the office. Tom Leonard amplified that feeling by resisting having a private corporate office as CEO. At first, he even kept a big sign outside of his office that read 'Wasted Space, Available for Manufacturing When Needed' as a way of showing how he felt about offices. Tom instilled that philosophy that every day was a dress-down day, unless it would make a customer feel uncomfortable in which case we would dress accordingly."

Vincent feels that the company is well-positioned for the future. "We have no debt, are not saddled with excess inventory, and have about $150 million in the bank," he notes. "We still have some military business—in fact, our Trans-Tech group has been on the Aegis program from the beginning—but we no longer pursue military business the way that we go after commercial markets. The company has a strong background in multiple-process technologies, such as GaAs HBTs, GaAs MESFETs, PHEMTs, and silicon discrete devices. For many years, we fought our own image with outside investors who considered us too broad across technologies, too complex to interest investors. But now, our breadth of technologies and products may provide the versatility to meet the changing, more integrated needs of emerging markets."

The manner in which Alpha located its Haverhill, MA facility demonstrates the high regard for its employees. "It took almost a year to find the Haverhill facility," says Vincent. "We looked at the demographics of where people lived, and who might be displaced if they were to shift working from the Woburn facility to the new facility. I don't think that we lost more than a few people when we moved to Haverhill. It was a way of being sensitive to the people who are employed here."

California Eastern Labs

California Eastern Laboratories (CEL) [Santa Clara, CA] is somewhat unique among microwave companies, being an exclusive supplier of RF and microwave semiconductors for Nippon Electric Co. (NEC) [Tokyo, Japan] in the US. The company literally began with a handshake. This classic story of honor and trust begins in 1959 when Al White started the company. He had left Sylvania convinced that there was a market in the US for vacuum tubes that were manufactured in Japan. And, according to Jerry Arden **(Fig. 2)**, the firm's current President, CEL actually served as the first manufacturer's representative for NEC in the US, even before NEC America.

"It was essentially a handshake between Al White and Dr. Kobyashi, the Chairman of NEC, in the airport that gave CEL the rights to sell NEC products in the United States. We have a contract now with NEC, but it is not as binding as that original handshake. We consider that handshake to be our legal document," says Arden, who joined the company in 1972. The relationship between the two companies has grown to the point where CEL is now responsible for all of NEC's microwave and laser products for North and South America.

NEC was one of the first companies to switch from germanium (Ge) to silicon (Si) for fabricating transistors, and one of the first companies to make true RF and microwave transistors. "In the early days, we were competing with Fairchild Semiconductor,

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CLV0815E	806	824	0.5-4.5	11	-113	-35	5.0	11
CLV0950E	865	1035	1-10	27	-114	-11	5.0	24
CLV0915A	902	928	0-4	17	-108	-30	3.0	10
CLV1085E	1050	1086	0.5-4.5	21	-112	-20	5.0	20
CLV1385E	1370	1400	0.5-4.5	18	-110	-20	5.0	20
CLV1550E	1500	1600	0.5-5.0	44	-106	-35	5.0	22
CLV2465E	2436	2496	1-4	26	-107	-20	5.0	25

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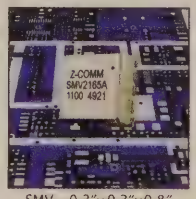
SMV0162A	125	200	0.7-8.3	12	-100	-6	5.0	36
SMV1570L	1540	1600	0.5-2.5	128	-90	-15	2.7	9
SMV2165A	2118	2218	0-3	148	-91	-10	3.3	16
SMV2390L	2290	2485	0-4	116	-90	-11	5.0	16
SMV2660L	2620	2700	0.5-4.5	90	-91	-17	5.0	21



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PLL0210A	200	230	100	0.50	-105	3.5±2.5	+5	25
PLL0930A	900	960	100	0.75	-101	3±2	+5	40
PLL1260A	1230	1290	1000	0.75	-102	1±2	+5	40
PLL1456A	1420	1490	1000	0.75	-103	1±2	+5	40
PLL2710A	2670	2740	1000	1.25	-98	1±4	+5	30

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3. Bob Traut has been involved in the development of RT/duroid® circuit materials and other products during a long and esteemed career at Rogers Corp.

Hewlett-Packard Co., and Motorola," recalls Arden. "The thing that really turned it for us was the GaAs FET. We had a chance to replace some GaAs FETs from domestic US suppliers because they were found to be unreliable. Our devices, on the other hand, were very reliable," notes Arden. "Once the customers were using our GaAs FETs, we were able to interest them in some of the other product lines. I remember the first device, the model V244, which was a one-micron GaAs FET. We then had the V388, which was a half-micron device, for lower-noise and higher-frequency operation."

During CEL's history, numerous individuals have contributed to its success, including Dick Lane, who ran CEL's microwave test laboratory. "Dick was really a 'Jack of all trades,' since he could design circuitry and also work with test equipment," says Arden. "He really pushed the envelope for what could be done in terms of microwave component measurements."

CEL currently employs approximately 175 people and performs joint development of products with NEC. "We have a joint development center with NEC here in the United States. We proposed it in 1984, and they agreed. We have about 20 engineers working on different projects," adds Arden.

Rogers Corp.

Rogers Corp. (Rogers, CT) began life in 1832 as the Rogers Papermaking Co., specializing in paper products that were manufactured on wet-cylinder paper machines. After 51 years at Rogers, Corporate Research Fellow Bob Traut remembers the paper products that preceded the microwave products. "When I started, we were a paperboard manufacturer. We were probably at about $3 to 4 million sales per year and trying to make it to $5 million," he says.

Bob Traut **(Fig. 3)** recalls the paper materials and manufacturing process: "A screen-covered cylindrical mold roll in a vat with circulating fiber slurry builds a paper ply as it rotates. It then transfers this to a continuous felt that builds a multiply sheet on a steel mandrel, where a burly worker removes it. In those days, the company's only connection to the electrical industry was paperboard insulation for use in large, oil-filled, power transformers. Earlier the company's major business was supplying cotton-rag, fiber-based paperboard called 'pressboard' for the area

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textile industry. As that business faded, the product found a critical role in electrical power distribution.

"I started a year out of high school as a quality-control tester, sampling wet paper plies and vat water from our four paper machines to control shrinkage related to degree of fiber hydration and pH for degree of acidity. Soon I transferred to the Research & Development department and began furthering my education both at Rogers and at the University of Connecticut."

He worked for the late Vice-President of R&D, Dick Berry, experimenting with many different fiber and polymer combinations. "If a fiber dispersed in water, we would look at it," Traut says. "In our R&D efforts, we used a pilot plant-scale paper machine with a wooden vat built before 1930."

4. The first RT/duroid® product, the 5870 material, was produced on a small-scale paper machine with a wooden vat.

Bob Traut was in high school when polytetrafluroethylene (PTFE) was discovered. "DuPont initially only sold the bulk resin in powder form not usable in our explorative work," he recalls. "Some DuPont people eventually came to Rogers with a problem with sheet PTFE flange seals for their processing equipment. They thought we might be able to help them. Their problem stemmed from the fact that when you clamp and heat PTFE, it creeps and leaks.

"So they provided PTFE as a stabilized aqueous colloidal polymer dispersion. We started experiments to develop PTFE/aluminum silicate ceramic fiber and PTFE/glass microfiber paperboards on our pilot paper machine. The first combination led to 5600 and 5650 creep-resistant grades that met DuPont's needs. The second led to 5870 and, later, 5880 grades that proved to be excellent microwave materials."

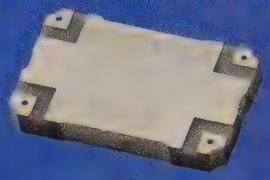
The 5870 material succeeded as waveguide windows for tracking and telemetry in the ablative nose cones of Jupiter rockets used for suborbital exploratory flights. "They were flying these rockets and losing re-entry cones in the Atlantic, because the earlier PTFE composites ablated away during flight" Traut says. "We machined waveguide windows from thick laminated blocks to lock into the nose cone. President Eisenhower appeared on television with the first recovered re-entry cone—the first to use 5870!

"We were contacted by microwave people from either Texas Instruments or Raytheon. They wanted to use our material for stripline circuitry, but none of us at Rogers knew what stripline was. They asked us if we could make our material with copper cladding. The initial elaborate process to add copper cladding to our 5870 material was horrible, requiring hot-air oven heating the material stack between thick sheets of asbestos composite to 730°F, well above the PTFE melting point. The fellow operating the process would wear an asbestos fire suit to open the oven, pull out the assembly, and put it in a platen press for clamping and cooling to make a copper clad laminate **(Fig. 4)**."

During his tenure at Rogers, Traut has learned more than a few tricks of the trade in developing not only new materials, but also the test methods to evaluate them and the software to orchestrate the test equipment.

He still looks back fondly at some of the challenges, including his project with process and materials for the Pershing II missile. "The Army had been developing the Pershing II missile as a tactical suborbital weapon with precise radar guidance during descent based on an electronic map of terrain contour of the target area," Traut notes. "The

90° ±1°
PHASE BALANCE



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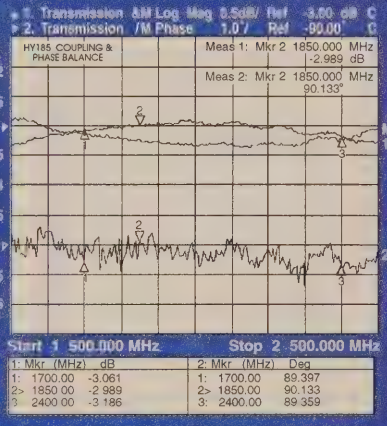

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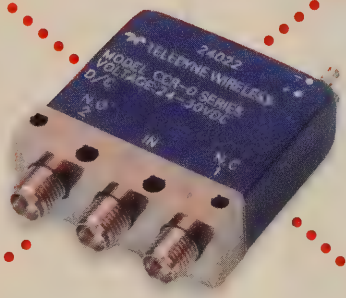

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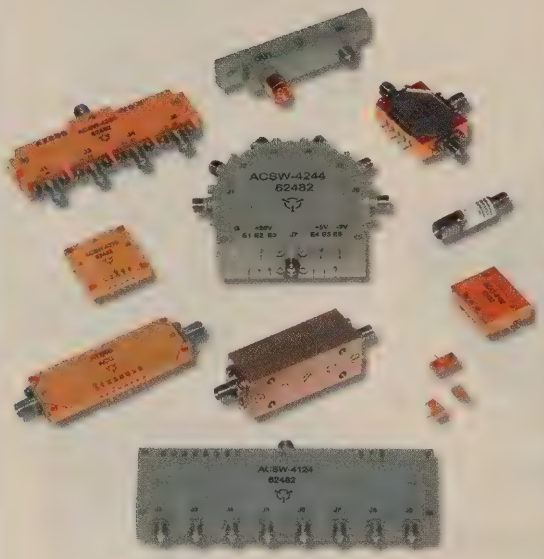

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original fused silica radome badly failed supersonic rain-erosion tests, and various PTFE-based ablative composites were little better.

"One day, I received a phone call from someone at Redstone Arsenal who had worked on the Jupiter project, asking for thick blocks of 5870. I just happened to have such a thick block in my office from experiments to fusion bond clamped stacks of sheet material. The dramatic first trial success on a three-stage rocket-sled test at White Sands Proving Ground excited the Army, and we ended up developing the Pershing II missile radome ablator. The prime contractor told us that this alone turned Pershing II into a strategic weapon."

The ablator appeared as the letter "i" in the word "Time" on this magazine's cover in the early 1980s.

"Some of our advances were by accident," admits Traut. "Someone mistakenly made a batch one day with only a small fraction of the specified fiber content. The surprise that this still ran in our process encouraged formulation of our 5880 grade with reduced fiber, lower dielectric loss, and good balance of other properties.

"At one point, our president had a vision of having every computer built with Rogers' materials. We were about 10 or 15 years too soon with that vision. But as computers go higher and higher in clock frequency, that vision starts to have some credibility."

> "Bob Traut's numerous patent awards attest to a brilliant career spanning more than five uninterrupted decades with a technology leader."

Today Bob Traut can be found at the famed Lurie Research and Development Center in Rogers, CT, where he works part-time. His numerous patent awards, which decorate the walls of his office, attest to a brilliant career spanning more than five uninterrupted decades with a technology leader in high-performance specialty materials.

Rogers Corp. provides specialty-materials solutions to customers worldwide, focusing on the wireless-communications and computer markets. Rogers operates manufacturing facilities in Connecticut, Arizona, and Illinois in the US and in Gent, Belgium. Sales offices are located in Japan, Hong Kong, Taiwan, Korea, and Singapore. Rogers has joint ventures in Japan with Inoac Corp., in Taiwan with Chang Chun Plastics, and in the US with Mitsui Chemicals and 3M Corp. The Advanced Circuit Materials Division at Rogers, located in Chandler, AZ, was recently formed by combining its flex-circuit materials and microwave-materials divisions. MRF

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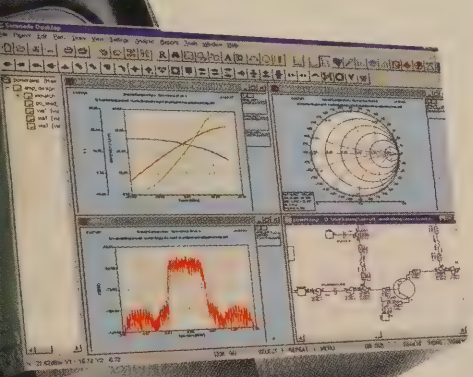

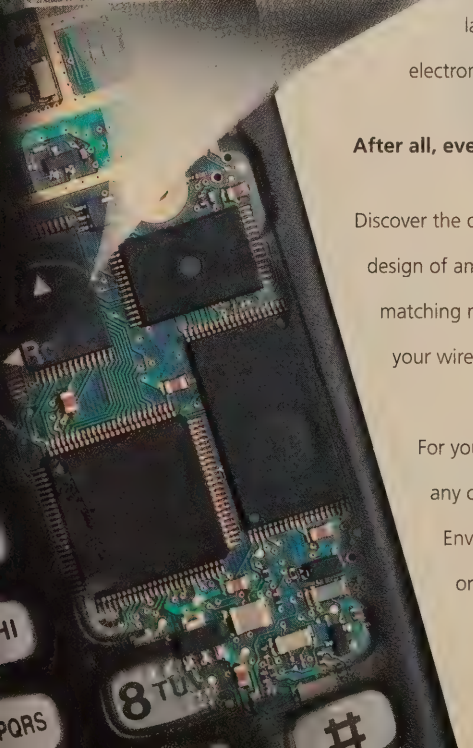

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Matrix targets Ka-band applications

WITH A FOCUS on Ka-band applications, the WR-51 waveguide switch block is designed to support redundancy systems for the new generation of multimedia communication satellites. The space-qualified matrix offers a design with three R switches that are integrated into one housing. DC characteristics include +25 to +35 VDC and operating current is 400 mA typical at 20°C. Equipped with a sequential latching actuator, the matrix switches in less than 500 ms. Frequency range is from 18.25 to 20.25 GHz and VSWR is 1.10:1 maximum. Insertion loss is 0.08 dB maximum per switch path and isolation is 50 dB minimum.

Dow-Key Microwave Corp., 4822 McGrath St., Ventura, CA 93003; (805) 650-0260, FAX: (805) 650-1734, Internet: www.dowkey.com.

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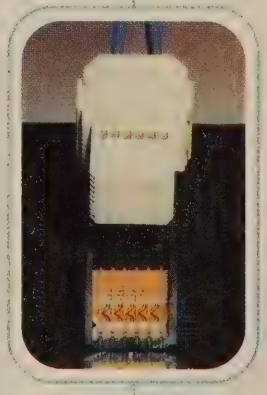

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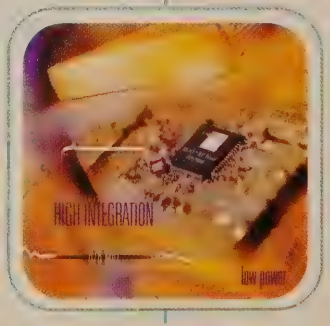

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Antenna Factor, 575 SE Ashley Pl., Grants Pass, OR 97526; (541) 956-0931, FAX: (541) 471-6251, Internet: www.antennafactor.com.

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LSI Logic Corp., 1551 McCarthy Blvd., Milpitas, CA 95035; (800) 574-4286, (866) 574-5741, Internet: www.lsilogic.com.

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War Clouds The Telecom Future

Before September 11, the telecommunications industry was in a serious economic slump, but some industry analysts guessed it was bottoming out. Then came the horrendous bombing of the World Trade Center and Pentagon on the second "day of infamy" in US history, which could mean that a telecom comeback has been pushed out to the indefinite future.

Analysts who were betting on a telecom turnaround predicated it on a broad economic recovery along with a proliferation of higher-bandwidth applications. Even before the terrorist attacks, the economy was teetering on recession (the official definition, not the reality). In the wake of the bombings, with large segments of American business virtually shut down for five days, an official recession will likely be called at the close of this economic quarter. Not only that, but the severe financial difficulties of businesses directly affected by the attack—airlines, aircraft manufacturing, insurance, retail trade—dwarf those of telecom and could push it to the back burner until the uncertainties of the present state of emergency become clearer.

On the bandwidth side, the general perception is a glut of this commodity, making the manufacture of more network equipment unnecessary. The excess results from a reduction in demand from end users, chief among them the failed Internet sites that no longer need the equipment. An often cited refrain that says users clamor for greater bandwidth may be accurate, but currently, that demand falls well short of the infrastructure supply built over the last few years. The excess capacity has a negative economical impact, reducing the revenue of carriers by deflating the prices of long-haul services.

Over the long run, however, Internet applications alone will drive up the demand for greater bandwidth. One reason is that the sizes of files downloaded by users from the global network has increased steadily—just with attachments to files, digital pictures, and audio. Future applications, such as video on demand, would require upgrades to the communications backbone even in the face of excess capacity.

With the nation in mourning, all economic and technical scenarios are on hold for the next few months. MRF

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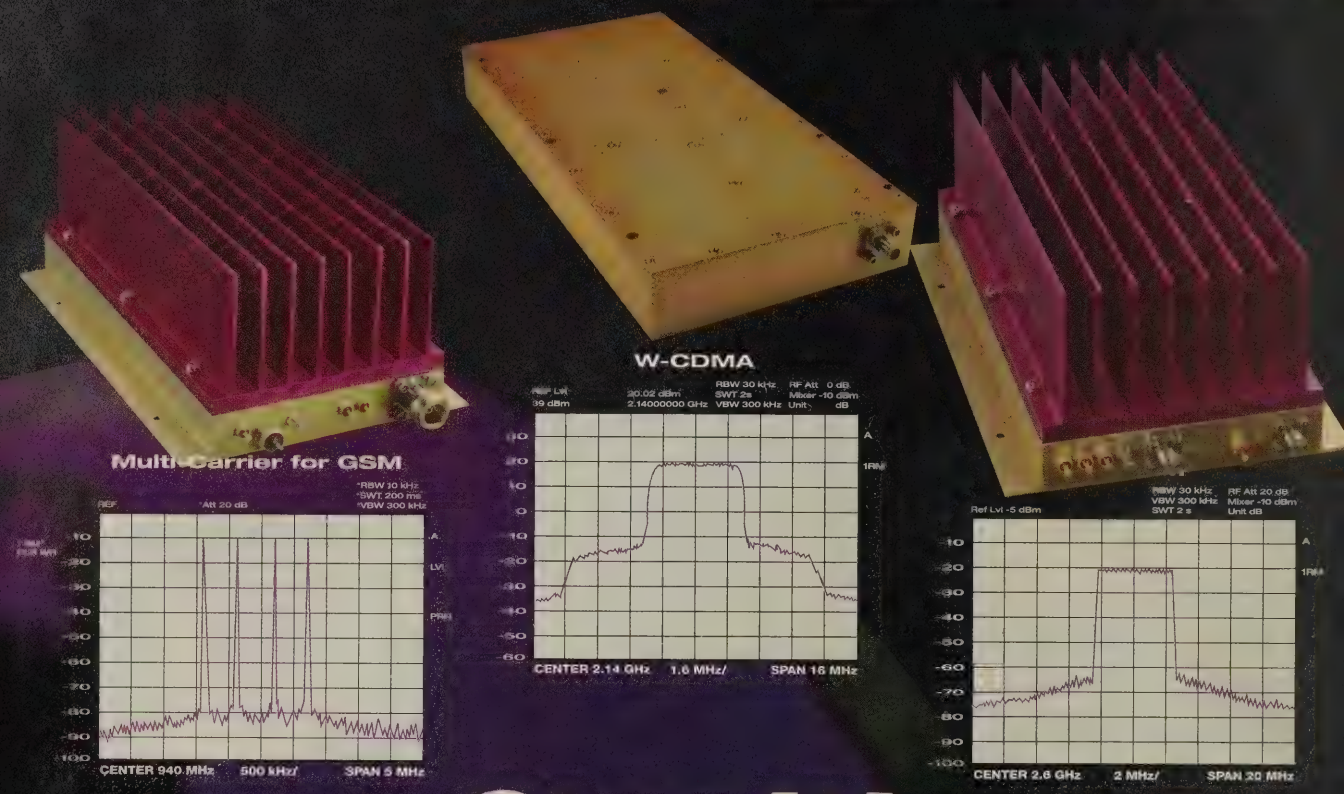

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CONTRACTS

Harris Corp.—Has begun shipments of its digital Falcon™ II radios to the US Army's 82nd Airborne Division—the largest parachute force in the free world.

RF Micro Devices—Has commenced production shipments of PAs to Sendo for use in their dual-band GSM (GSM-900/1800) mobile phone, the S200. The RFMD® PAs are GaAs HBT MMICs that provide high efficiency in small, thermally enhanced, cost-effective packages.

Andrew Corp.—Was awarded a contract from Hong Kong's MTR Corp. to upgrade the RF signal underground. This will enable better reception and improved voice quality for commuters using mobile-communications devices within the MTR's trains and stations.

TestMart—Has been awarded a contract with an estimated value of $600 million by the Department of Defense (DoD) Naval Inventory and Control Point (NAVICP). Under terms of the five-year agreement, TestMart will develop and operate a private e-commerce marketplace for the General Purpose Electronic Test Equipment (GPETE) division. NAVICP estimates that the contract will reduce procurement costs by more than $360 million.

Verizon Wireless—Has purchased the Air Access C2K CDMA Network Emulator product and the C2K-ATS Automatic Test System product from Spirent Communications. Terms of the purchase were undisclosed.

Motorola's Global Telecom Solutions Sector—Has signed a $47 million network-infrastructure expansion deal with Maroc Telecom, a Morocco Mobile Operator. Motorola is contracted to provide the new generation of Horizon macro GSM-900 megahertz base stations for the popular tourist region of Marrakech and Agadir in the southern part of Morocco, as well as in the northern cities of Tangier, Tetouan, Fes, Meknes, Casablanca, and El-jadida. The deal, which is an extension of Motorola's existing business with Maroc Telecom, means that Motorola continues as a major base-station supplier to Maroc Telecom's GSM mobile network.

FRESH STARTS

Ipricot North America—Will make its line of satellite broadband-communication equipment commercially available through BitCentral's online website. Bitcentral.com is the only full-service e-commerce website for buyers and sellers of satellite broadband technology.

Accelerated Technology, Inc. (ATI) and Skelmir, LLC—Have signed an agreement to expand their relationship from a partnership to a reseller agreement. According to the agreement, ATI will resell Skelmir's Cee-J virtual machine out of the box to Nucleus developers who are building Java-enabled devices.

Galtronics—Announced the completion of its state-of-the-art wireless design center in Phoenix, AZ. The Phoenix design center is part of Galtronics' corporate headquarters and includes an anechoic chamber, SAR testing equipment, network analyzers, machinery for prototype molds and components, CAD/CAM and FEA software, and plans for a reliability lab. An NSI anechoic chamber with near-field testing capabilities of 800 MHz to 18 GHz, as well as a far-field testing range, are featured at the Phoenix site. Galtronics also has state-of-the-art design centers in Tiberias, Israel and Tianjin, China.

SnapTrack, Inc. and TechnoCom Corp.—Announced a joint effort to provide integration and deployment support for E911 emergency cell-phone location and location-based commercial services for Qualcomm's SnapTrack-enhanced, gpsOne™-based wireless-location systems. The agreement between the two companies establishes TechnoCom as the preferred field-test, engineering-services, and integration contractor for carriers and SnapTrack OEMs for deployment services, trial management, and implementation of gpsOne-based wireless-location systems in the US.

Avnet's RF Design Services Bluetooth Competency Center—Has introduced Avnet Bluekit™. The kit features Philips Semiconductors' PCF87750 Bluetooth baseband controller, which supports point-to-multipoint Bluetooth connections and Bluetooth v1.1 with a BGB100 radio module. The Avnet Bluekit is designed to enable rapid adoption of Bluetooth technology by application developers into their end products. The solution is an integrated device with Flash memory and codec built in.

Gabriel—Has signed Entest, Inc. of Carrollton, TX to represent Gabriel's Wireless Broadband point-to-point and point-to-multipoint antenna products, transmission-line systems, and pressurization equipment in the South Central region of the US. This agreement enables Entest, Inc. to represent Gabriel in the states of Texas, Oklahoma, Arkansas, Louisiana, and Mississippi.

Agilent Technologies, Inc. and Frontier Electronic Systems Corp. (FES)—Announced that FES has joined Agilent's Premier Solution Partner (PSP) Program to provide customers with a range of commercial and military applications in aerospace and defense.

Trimble and XEMICS—Through a joint sales agreement, the companies have announced that InnoLabs Corp. will be adding GPS functionality to its personal-computing tablet. InnoLabs will be purchasing the XEMICS XE1610 GPS chip set. The chip set, which is based on Trimble's FirstGPS™ technology, allows high-volume manufacturers of consumer products to add GPS location capability with minimal impact on device size or battery life. The chip set from XEMICS is the smallest and lowest-power implementation of Trimble's FirstGPS technology to date. It is specifically designed for power-sensitive, wireless enabled, mobile information devices such as laptops, PDAs, digital cameras, and even wristwatches. MRF

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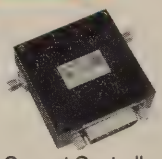

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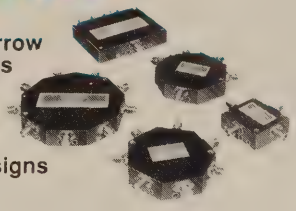

SPST thru SP8T and Transfer type models are offered and all switches are low loss with isolation up to 100dB. Reflective and non-reflective models are available along with TTL compatible logic inputs. Switching speeds are 1μsec.—30nsec. and SMA connectors are standard. Custom designs including special logic inputs, voltages, connectors and package styles are available. All switches meet MIL-E-5400

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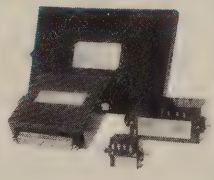

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Rohde & Schwarz Appoints Catherine To Post

PHILIPPE CATHERINE has been named to the post of managing director of sales in Europe for Rohde & Schwarz. Most recently, he was managing director of Rohde & Schwarz France S.A. and president of the board of directors of Rohde & Schwarz España.

OSE Systems—W. VIRGINIA WALKER to CFO and executive vice president of finance and administration for OSE Systems worldwide; formerly CFO and vice president of finance and administration at Vyou, Inc.

Gould Fiber Optics—RICHARD A. KEENEY to manager of the New Freedom, PA plant; formerly vice president and program director for the Weather and Air Traffic Systems division at Systems Management, Inc.

Uniroyal Technology Corp.—DAVID C. MILLER, PH.D. to president of the Optoelectronics business segment; formerly president of the Airtron Division at Litton Systems.

appli-tec—DAVID W. WILCZYSKI to operations manager of the encapsulation and potting division; formerly engineering manager.

The Phoenix Co.—TERRENCE J. MADIGAN to vice president; formerly general manager.

Endwave Corp.—KENNETH DER to vice president of finance and corporate controller; formerly corporate controller. Also, BRYSON WALLACE to vice president of engineering; formerly manager of the engineering teams, product development, and customer engineering programs.

REMEC, Inc.—BRUCE ANDERSON to executive vice president for manufacturing operations; formerly vice president/general manager for DMC Stratex.

Palomar Technologies—BURKHARD HABBE to European sales manager, responsible for sales in Spain, France, Scandinavia, Eastern Europe, and the Middle East; formerly sold die and wire bonders in the major part of Europe as area sales manager for F&K Delvotec.

Also, MARTIN BRANDL to regional sales manager for the German-speaking countries, Benelux, and Italy; formerly sold die bonders at Datacon. In addition, DAVID MUDD to regional sales manager for the United Kingdom; formerly project leader for photonic packaging automation at Nortel Networks.

Antenna Specialists—MICHAEL UXA to national sales manager for the Consumer Products Group; formerly regional sales manager for the professional and cellular product lines.

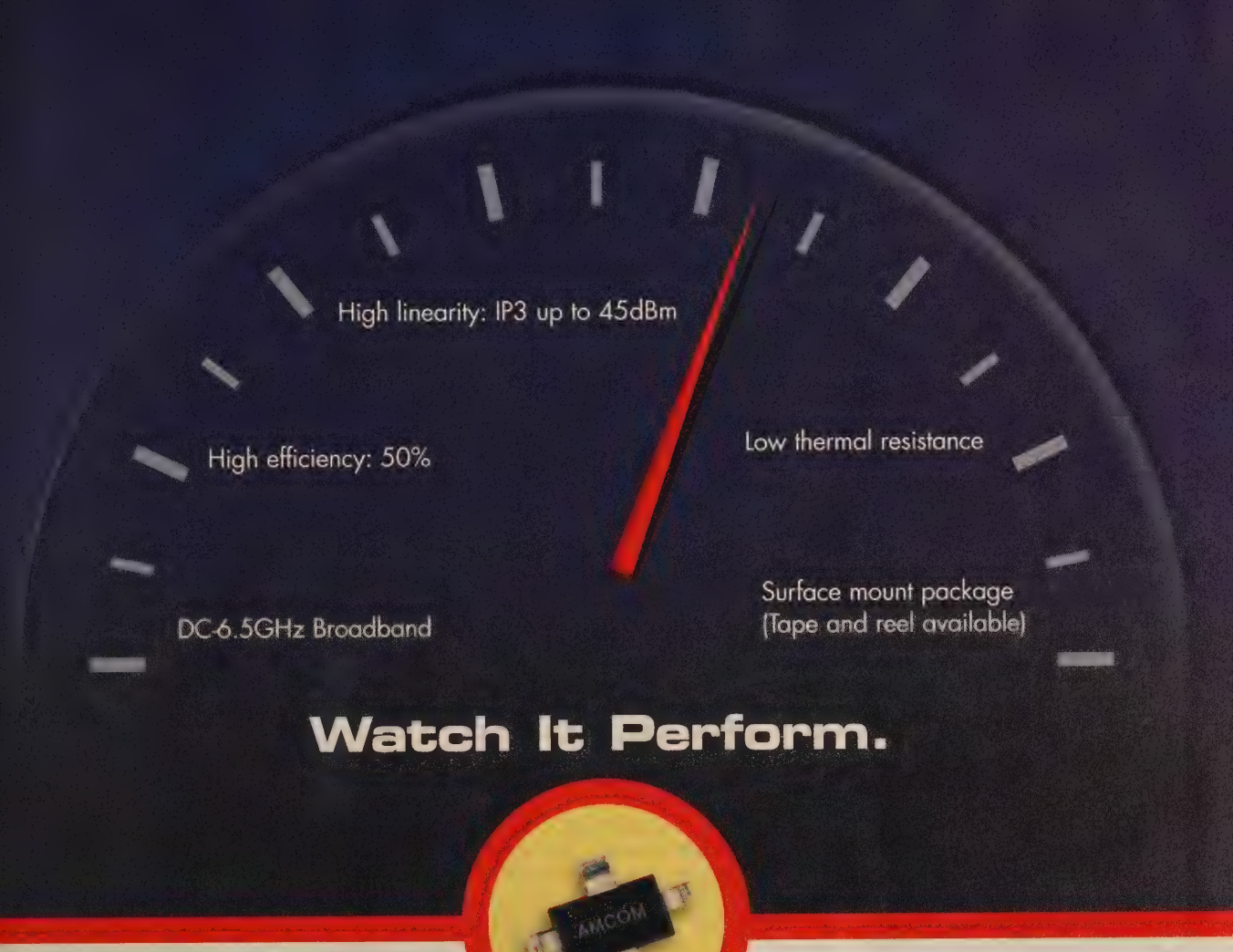
Narda—GERARD SPINELLI to sales engineer; formerly account manager for East Coast Microwave.

Vitesse Semiconductor Corp.—RICHARD RIKER to vice president of marketing and sales; formerly vice president of sales and business development at Stargen.

Hybrid Networks, Inc.—THOMAS BISSETT to vice president of engineering; formerly vice president of manufacturing and support.

U.S. Cellular—SHERRY DAVENPORT to director of segment marketing; formerly held the title of manager of business marketing.

Kathrein, Inc.—DAVID LITTLE to East Coast regional sales manager; formerly employed in antenna application at Allgon and dB Product. MRF



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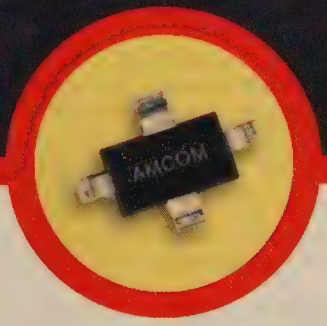

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AM024MX-QG	DC-6.5	13dB	29dBm	41dBm	46%
AM036MX-QG	DC-6.5	12dB	31dBm	43dBm	46%
AM048MX-QG	DC-6.5	11.5dB	32dBm	44dBm	46%
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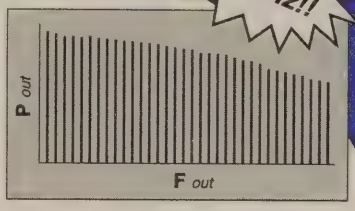

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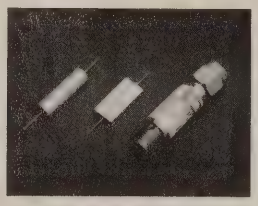

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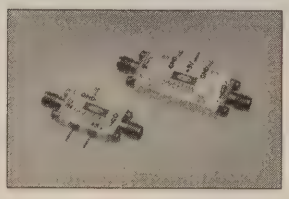

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▸ MEETINGS

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Fayetteville, AR 72701
e-mail: apm2@engr.uark.edu

Asia-Pacific Optical and Wireless Communications Conference and Exhibit 2001
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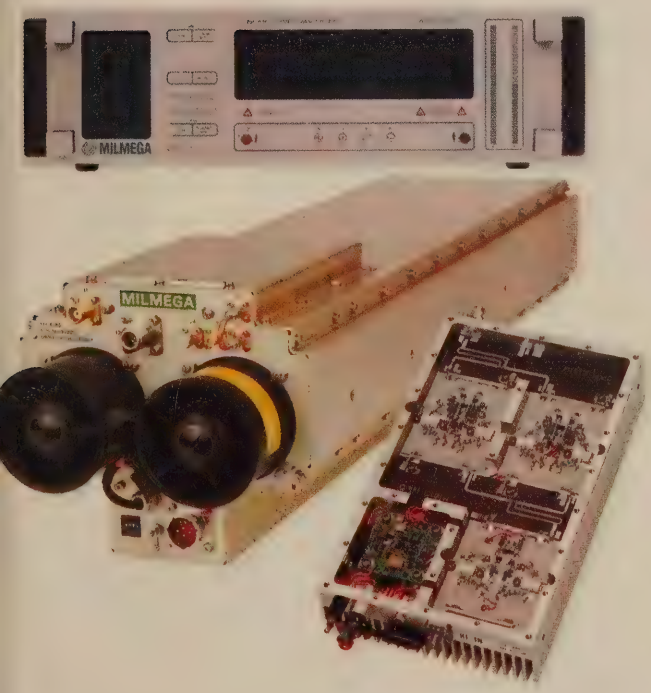

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Fiber-Radio System Serves Indoor 60-GHz ATM Network

HIGH-DATA-RATE NETWORKS permit the bidirectional transfer of large files, such as multimedia and video, in office environments. To address the need for speed in these networks, S. Dupont and a team of researchers from the Universite des Sciences et Technologies de Lille (Villeneuve d'Ascq, France) have developed a 60-GHz fiber-radio communication system for indoor ATM networks. The system, which is transparent to a wide range of digital modulation schemes, uses two microwave signals that are transmitted through a 10-km optical fiber. The first 1-GHz subcarrier supports high-data-rate information, while the second 7-GHz reference signal features the signal purity needed to limit degradation of radio-communications quality. Each signal modulates an optical laser. Following optical transmission, the 7-GHz signal is demultiplexed and photodetected, then frequency multiplied to 56 GHz to act as the pump signal for upconversion and downconversion. The system features MMIC T/R modules based on PHEMT technology. To evaluate the system, a full-duplex QPSK radio link was tested with a data rate to 100 Mb/s at communication distances from 4 to 12 m with good results. The system, which employs circular polarization but does not require additional equalization, was also tested over longer distances, with data rates to 22 Mb/s possible. For more information on this innovative communication system, see "60 GHz Fiber-Radio Communication System For Indoor ATM Network," *Microwave and Optical Technology Letters,* September 5, 2001, Vol. 30, No. 5, pp. 307-310.

Continuous-Time ADC Clocks At Rates To 18 GHz

FAST ADCS HAVE long been on the "shopping list" of military systems designers needing to perform instantaneous signal processing on communications and radar signals. Hope for an all-digital Rx front end can be found in the work of Shrinivasan Jaganathan and associates from the University of California at Santa Barbara (Santa Barbara, CA), who have reported a continuous-time, second-order Σ-Δ ADC with clock rate to 18 GHz. The converter features an ENOB on the order of 10 at 200 MHz, with 6.8-ENOB resolution at 500 MHz. The ADC is fabricated with InAlAs/InGaAs transferred-substrate HBT technology, using transistors with current-gain cutoff frequency of approximately 190 GHz and maximum frequency of oscillation of 200 GHz. The ADC IC, which was tested by on-wafer probing with 40-GHz probe cards, consists of two transconductance cells, each followed by a passive integrator. The quantizer is a master-slave flip-flop, while the feedback DAC is a current-steering differential amplifier. An error signal is generated by current-summing at the output nodes of the first transconductance cell. The ADC was tested at a clock rate of 18 GHz using signal bandwidths from 150 to 900 MHz and performing analysis on a high-speed sampling oscilloscope. The modulator achieved a two-tone SNR of 48, 42, and 33 dB, respectively, for input signals at 150, 500, and 990 MHz, respectively. For more information, see "An 18-GHz Continuous-Time Σ-Δ Analog-Digital Converter Implemented in InP-Transferred Substrate HBT Technology," *IEEE Journal of Solid-State Circuits,* September 2001, Vol. 36, No. 9, pp. 1343-1350.

Three-Dimensional MMICs Cut Costs For Integrated Millimeter-Wave Circuits

DEMAND FOR LOW-COST millimeter-wave circuits has been steadily increasing with the potential of markets for automotive, high-data-rate WLANs, and LMDS. Three-dimensional MMIC technology offers the high levels of integration and manufacturing simplicity needed to achieve the cost targets of these applications, and impressive work on 3D MMIC technology has been reported by Kenijiro Nishikawa and associates at NTT Network Innovation Laboratories (Atsugi, Japan), NTT DoCoMo, Inc. (Yokosuka, Japan), and NTT Photonics Laboratory (Atsugi, Japan). Working with device technology from Matsushita/Panasonic (Tokyo, Japan), the researchers applied NTT interconnection technology to fabricate several 0.15-μm PHEMT-based MMICs, including a V-band balanced doubler and a V-band amplifier. The doubler, which measures 1.13 $\times$ 1.05 mm, achieves 0-dBm output power and more than 35-dB fundamental signal suppression from 31.5 to 37 GHz. The amplifier, which measures 1.24 $\times$ 0.43 mm, achieves approximately 10-dB gain from 50 to 70 GHz. The doubler is designed for use at +2 VDC, while the amplifier operates at +3 VDC. For more information, see "Three-Dimensional MMIC Technology for Low-Cost Millimeter-Wave MMICs," *IEEE Journal of Solid-State Circuits,* September 2001, Vol. 36, No. 9, pp. 1351-1359.

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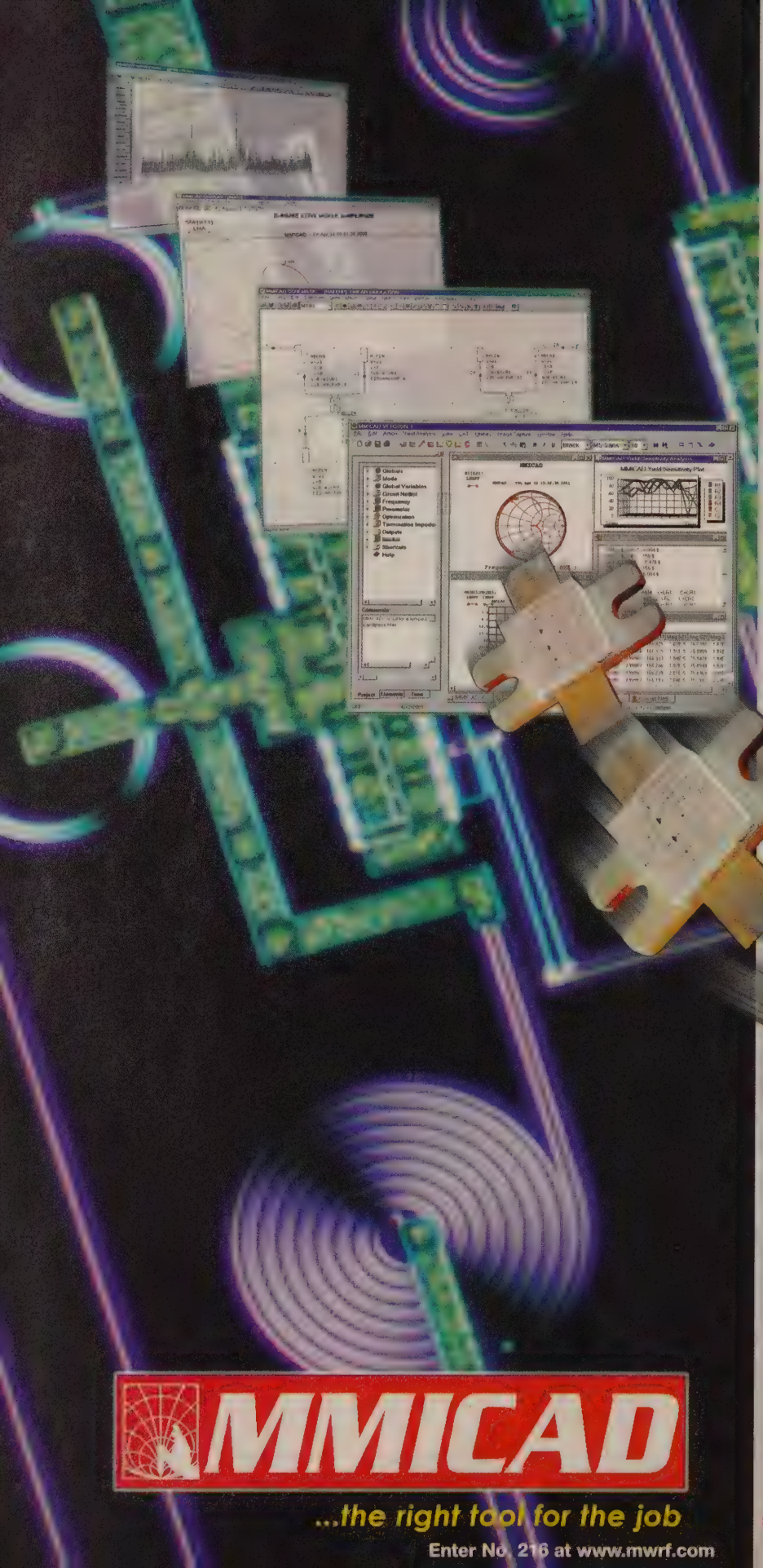

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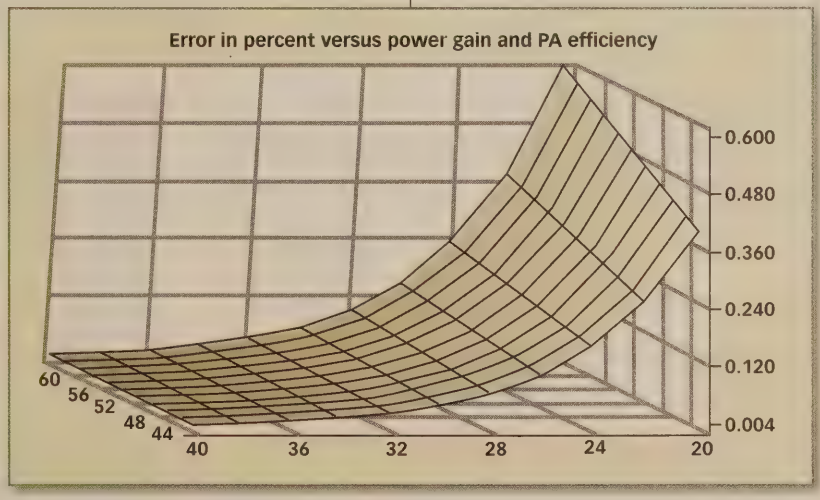
Mobile phones are increasing in popularity and use. They have been transformed from a tool of the business elite to a necessity for people from all walks of life. Just as families have come to purchase more than one automobile, they are also buying more than one mobile phone. In this transition, the cost of mobile calls has decreased, while usage has increased dramatically. Mobile phones are also evolving from something for emergency use to a product that stores address books, calendars, journal entries, and task lists. These mobile phones are now mobile terminals used for web browsing and e-mail.

This evolution has imposed new demands on the batteries and circuitry in the mobile terminal. The new requirements for talk time have brought about great pressure to provide very-low power-consumption circuitry. While 30 minutes of talk time was once standard, today's mobile phones have more than six hours. The power-amplifier (PA) circuitry imposes the greatest current draw on the battery and should be a key focus of power conservation. This article will review the basic assumptions about efficiency, the losses associated with different power-



1. In these curves of efficiency error as a function of gain and approximated efficiency, the x-axis represents power gain from 20 to 40 dB, the y-axis is efficiency, and the z-axis is percent error.

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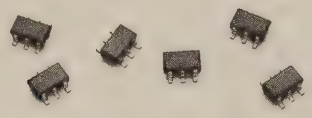

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UPC2747TB	Low Noise - 1900 MHz	1800	12	-7	3.3	5	900
UPC2748TB	Low Noise - 1900 MHz	1500	19	-3.5	2.8	6	900
UPC2749TB*	Low Noise - 900 MHz	2900	16	-6	4	6	1900
UPC8119/20T	AGC Driver Amp	1900	14	3	9	11	900

*For GPS RF amplifier applications

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UPC8151TB	1900	12	4	—	4.5	900	S06
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*Specified for operation at frequencies to 2.4 GHz

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UPC2763TB	2400	20	8	5	27	1900	S06
UPC2771TB	2400	20	13	5	35	900	S06
NEW UPC8181TB	4000	19	9.5	4.5	23	900	S06
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Part Number	3dB BW (MHz)	Gain (dB)	P_{SAT} (dBm)	NF (dB)	I_{CC} (mA)	Test Freq (MHz)	Package Style
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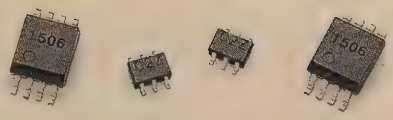
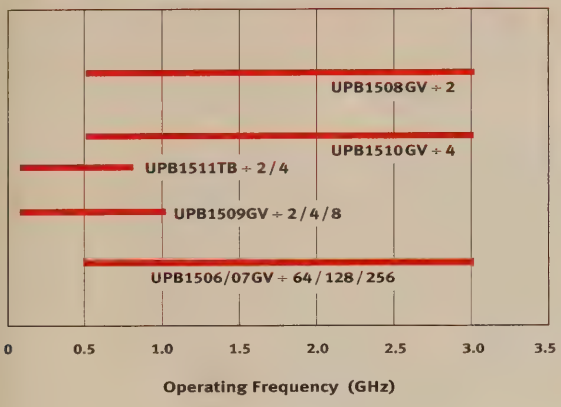
UPC2791TB	1900	12	4	5.5	17	500	S06
UPC2792TB	1200	22	5	4.5	19	500	S06
UPC2708TB	2900	15	10	6.5	26	1000	S06
UPC2709TB	2300	23	11.5	5.0	25	1000	S06
UPC2710TB	1000	33	13.5	3.5	22	500	S06
UPC2711TB	2900	13	1	5	12	1000	S06
UPC2712TB	2600	20	3	4.5	12	1000	S06
UPC2713T	1200	29	7	3.2	12	500	T06
UPC2776TB	2700	23	8.5	6	25	1000	S06
NEW UPC3215TB	2900	20.5	3.5	2.3	14	1500	S06



5V Silicon Prescalers



3V Silicon Downconverters...

Part Number	RF Freq (MHz)	IF Freq (MHz)	CG (dB)	IIP_3 (dBm)	NF (dB)	I_{CC} (mA)	Test Freq (MHz)
UPC2756TB	100-2000	10-300	14	-14	13	5.9	900
UPC2757TB	100-2000	20-300	13	-15	13	5.6	800
UPC2758TB	100-2000	20-300	17	-14	13	11	800
UPC8112TB	800-2000	100-300	13	-10	11	8.5	900

... and Upconverters

Part Number	RF Freq (MHz)	IF Freq (MHz)	CG (dB)	IIP_3 (dBm)	NF (dB)	I_{CC} (mA)	Test Freq (MHz)
UPC8106TB	100-2000	50-400	9	-4.5	8.5	9	900
UPC8163TB	100-2000	50-400	9	+0.5	12.5	16.5	900
NEW UPC8172TB	800-2500	50-400	9.5	-2	9.5	9	900
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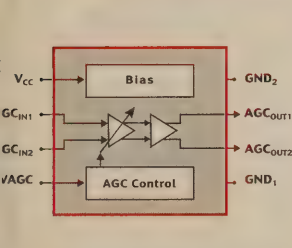

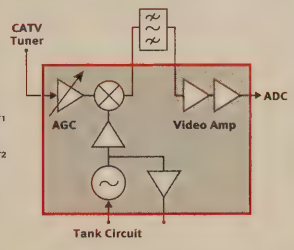

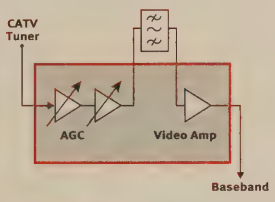

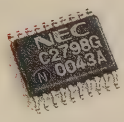

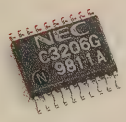

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control architectures, and will propose a new definition for PA efficiency.

For the purpose of this article, it is important to define efficiency. The fundamental definition for efficiency (Eff) is power out divided by the sum of the DC power in Eq. 1:

$$Eff = \frac{P_{out}}{\sum_{m} P_{(DC_m)}} \qquad (1)$$

The total sum of generated and dissipated RF power is shown in Eq. 2, where P_n denotes power generators and dissipaters and P_{in} denotes the input power.

$$P_0 + P_1 + P_2 + \ldots\ldots$$
$$P_{n-1} + P_n - P_{in} - P_{out} = 0 \qquad (2)$$

And rearranging Eq. 2 in summation form:

$$\sum_{n} P_n - P_{in} - P_{out} = 0 \qquad (3)$$

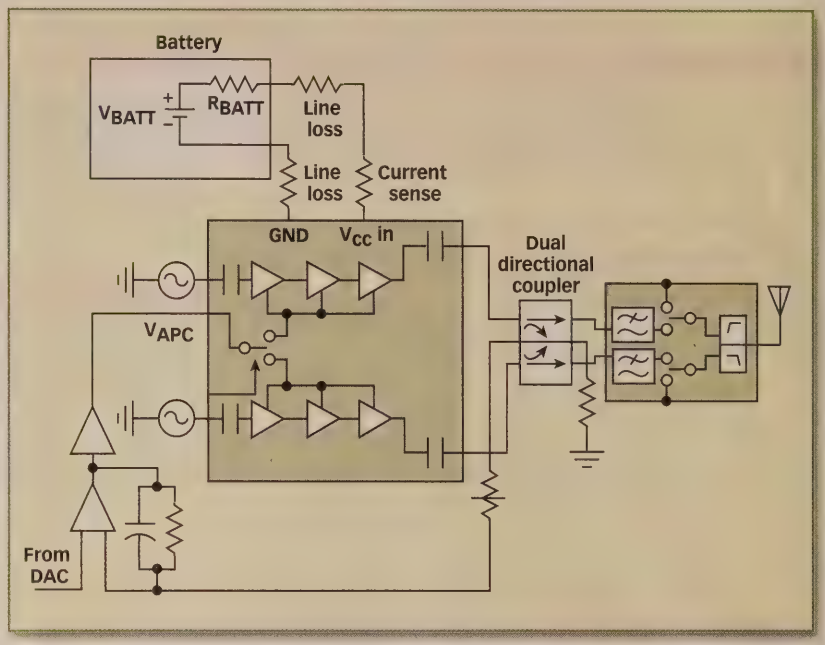
Equation 3 is solved for P_{OUT} divided by the summation of all DC input power, providing the equation for efficiency η_{sys} that is given in Eq. 4:

$$\eta_{sys} = \frac{\sum_{n} P_n - P_{in}}{\sum_{m} P_m} \qquad (4)$$

To more accurately state the efficiency of the PA, PA designers use power-added efficiency (PAE), which considers the efficiency of the power gain in the amplifier (Eq. 5):

$$PAE = \frac{P_{out} - P_{in}}{P_{DC}} \qquad (5)$$

For multistage high-power amplifiers, it is possible to approximate PAE by using Eq. 6. The error that is in this assumption is negligible for amplifiers that have more than 28 dB of gain and less than 60-percent efficiency. In the case of a PA with 28 dB of gain and 60-percent efficiency, the error in this assumption is less than 0.1 percent. Equations 5 and 6 can be combined in order to quantify the error where the gain is expressed as linear gain will be shown. **Figure 1** shows the approximation error as a function of gain and efficiency. Due to the negligible error and for simplicity of explanation, many of the following efficiency discussions will approximate PAE by using Eq. 6.



2. The DC supply-loss model of a GSM Tx shows the battery loss, and supply- and ground-resistance losses. RF losses are present in the directional coupler, filters, switch, and antenna.

$$\eta = \frac{P_{out}}{P_{DC}} \qquad (6)$$

The error associated with Eq. 6 is calculated as:

$$Error = \frac{P_{OUT}}{P_{DC}} - \frac{P_{OUT} - P_{IN}}{P_{DC}} \qquad (7)$$

P_{DC} can be related to efficiency and P_{OUT} to yield:

$$Error = \frac{P_{OUT}}{\frac{P_{OUT}}{\eta}} - \frac{P_{OUT} - P_{IN}}{\frac{P_{OUT}}{\eta}} \qquad (8)$$

Realizing that linear gain is defined as output power divided by input power reduces Eq. 8 to Eq. 9:

$$Error = \frac{\eta}{gain} \qquad (9)$$

Notice that Eq. 9 is independent of output power. The efficiency of PAs is often discussed in the context of Eqs. 5 and 6, but it is important to discuss these efficiencies in the context of system performance. In considering system performances, DC and post PA losses can make an otherwise efficient PA effectively inefficient. **Figure 2** illustrates many components in a Global

Table 1: GSM margin analysis

COMPONENT	NOMINAL LOSS	WORST-CASE LOSS (dB)	REQUIRED MARGIN (dB)
Directional coupler	0.3 dB	0.5	0.2
Antenna switch	1.1 dB	1.3	0.2
Mismatch loss at antenna	0.1 dB	0.5	0.4
Margin to prevent AM-to-AM conversion			0.5
Trace losses	0.2		
Total margin			1.3

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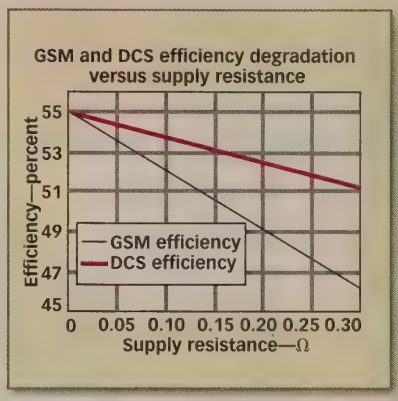
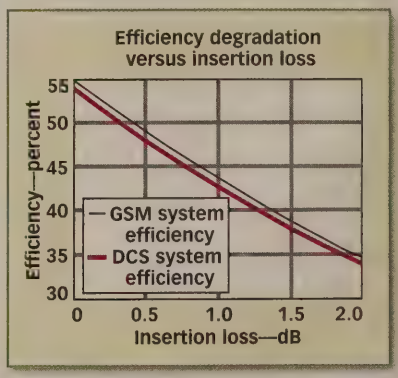


3. Efficiency falls off as supply losses grow larger as shown by these curves of increasing supply resistance versus efficiency.

System for Mobile Communications (GSM) system with a loss model for the DC supply. Three main DC losses are illustrated in the diagram. The first is the internal resistance of the battery which, in the case of a lithium-ion (Li-ion) battery, is approximately 50 mΩ. Another DC loss is the resistance of the supply and ground lines to and from the PA. Although these may be small, they affect system efficiency. Some systems control output power by sensing the PA supply current. This also adds a DC loss to the system. These losses are very important to track, since those associated with current-sensing power control typically reduce system efficiency by more than 3 percent. It is easy to reach the 100-mΩ level through contact resistance and line loss. Although it is possible to more accurately model



4. Efficiency degrades as insertion loss increases as these curves show for either a GSM or DCS PA.

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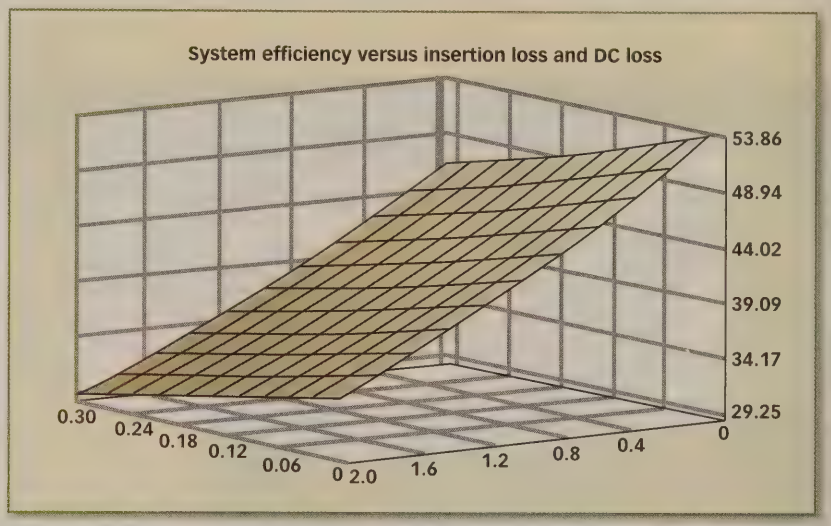
the DC losses in a transmitter (Tx), the scope of this discussion will be limited to the aforementioned losses.

The RF losses are also important to consider when analyzing system performance. Figure 2 shows two categories of RF losses. The first is the post PA loss. This consists of losses in the directional coupler, the filters, switch, and antenna. Since the system blocks are not truly 50-Ω interfaces, there are also mismatch losses between each block.

The second, less obvious RF loss is known as effective loss and is the result of margin that is designed into the system. For the system in Fig. 2, the effective losses consist of the power-control margin and component-variation margin. GSM power-control systems usually allow 0.5 to 1.0 dB of margin to prevent the onset of amplitude-modulation-to-amplitude-modulation (AM-to-AM) conversion that results in switching spectrum failures. The variation margins are comprised of frequency variation and loss variation in the directional coupler, low-pass filter, and antenna switch **(Table 1)**. These variations can result in 1.3 dB of margin added to the system. When a GSM PA operates at 1.3 dB below maximum output power, the efficiency is reduced by approximately 5 percent.



5. System-efficiency degradation from supply resistance and insertion loss is shown here. The x-axis is insertion loss from 0 to 2 dB, the y-axis is DC loss, and the z-axis represents efficiency in percent.

After considering the margin analysis, it is clear that when more components are added to a system, the required margin to maintain high yield also increases **(Table 2)**. The best solution for high yield and high efficiency is to minimize the number of components in the system. Yield improvements that are associated with component reduction will be discussed in greater detail in another article of this series.

Table 2: DCS margin analysis

COMPONENT	NOMINAL LOSS	WORST-CASE LOSS (dB)	REQUIRED MARGIN (dB)
Directional coupler	0.4 dB	0.6	0.2
Antenna switch/filters	1.5 dB	1.7	0.2
Mismatch loss at antenna	0.1 dB	0.5	0.4
Margin to prevent AM-to-AM conversion			0.5
Trace losses	0.4		
Total margin			1.3

Table 3: Sources of DC loss in a mobile-phone system

COMPONENT	RESISTANCE (mΩ)	NOTES
Battery	50	Internal resistance of the battery
Battery contact	30	At each contact
Supply feed track	75	60-mm line, 1 mm wide
Current-sense resistor	82	For systems using current-sense power control
Total	180 to 262	

In systems that operate between +3 and +4 VDC, small DC losses contribute to a significant drop in system efficiency. For these systems, a 200-mV drop equates to a 5-to-7-percent drop in supply voltage. Since power is a function of voltage squared, it is important to consider these losses. **Figure 3** shows the drop in system efficiency as the supply loss increases. Note that for increasing values of supply resistance, the system efficiency drops to 5 percent with only 200-mΩ resistance between the battery and the PA. Although 200 mΩ seems high, it is easy to reach. Using a Li-ion battery and very good supply lines, the DC losses will approximately be 200 mΩ.

The losses in the supply line are calculated by multiplying the cross-sectional area of the line by the length of the line and the resistivity of copper (Cu). Equation 10 shows this relationship where ρ is the resistivity of Cu ($17 \times 10^{-9}\ \Omega - m$), L is the length, A is the area, and $\alpha = 0.0039$ is the temperature coefficient for Cu. A supply feed of approximately 60 mm, including vias, has approximately 75-mΩ resistance. In addition to the line loss, there is also contact resistance between the battery and the supply line. Many contact connections have approximately 30-mΩ contact resistance. The typical

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DC losses are shown in **Table 3**.

$$R_{line} = \rho_0\left[1 + \alpha(T - T_0)\right]\frac{L}{A} \quad (10)$$

To understand the impact on system efficiency, Eq. 6 can be modified by eliminating P_{out}:

$$\eta_{sys} = \frac{\eta V_{pa} I_{cc}}{V_{batt} I_{cc}} \quad (11)$$

V_{pa} is the voltage that is applied to the PA and accounts for the voltage drop across the supply losses.

$$V_{pa} = V_{batt} - I_{cc}R \quad (12)$$

Equation 13 indicates that system efficiency is a function of PA efficiency, supply losses, and the supply.

$$\eta_{sys} = \eta\left(1 - R\frac{I_{cc}}{V_{batt}}\right) \quad (13)$$

By analyzing the supply conditions, the supply current divided by supply voltage can be related to load impedance through the load-line equation.

$$\eta_{sys} = \eta - \frac{R}{2R_{load}} \quad (14)$$

The dependence on load impedance can be understood in the context that the load line and output power are defined by the load impedance. Lower load impedances result in higher output powers as well as higher currents and lower system efficiencies. A GSM PA has a load impedance of 2 Ω or less, whereas a digital-communications-services (DCS) PA has a load impedance of approximately 4 Ω.

After considering the dependency system efficiency has on supply losses, more effort should be focused on lowering supply losses rather than higher-efficiency PAs.

Now that the ohmic losses are understood, it is important to explore the effect of insertion loss on system efficiency. Due to the maturity of the market, antenna switches, directional couplers, lowpass filters, and other components have very similar performance regardless of manufacturer. Considering PA efficiency solely is technically incorrect and provides an inaccurate

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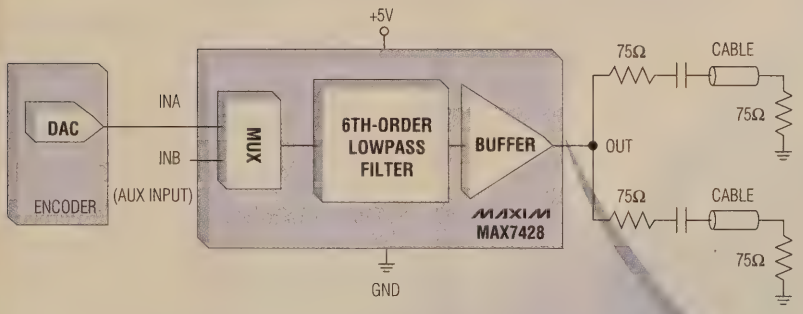

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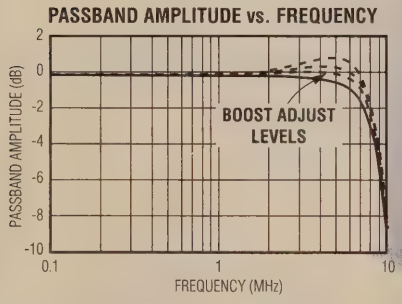

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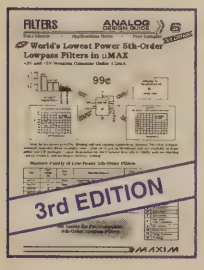

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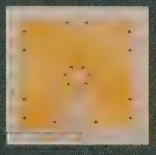

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picture of the system performance. "Efficiency" should also embrace the total loss in the Tx and be discussed in terms of system efficiency.

If the DC losses are omitted in Eq. 4, the system efficiency equation can be simplified to:

$$\eta_{sys} = \frac{P_{pa} - P_{in} - P_{loss}}{V_{dc} I_{dc}} \qquad (15)$$

Equation 15 is plotted as a function of P_{loss} in **Fig. 4**. This demonstrates the effect insertion loss has on the efficiency of a GSM PA with 55-percent efficiency and a DCS PA with 54-percent efficiency. As Eq. 15 shows, the slope of the efficiency degradation is equal regardless of the initial efficiency.

Unified System Efficiency

System efficiency is always a function of the degradations due to the DC loss and RF insertion loss. So far, the impact on efficiency has been described as a function of either ohmic loss or insertion loss but has not been described in a single equation. A unifying equation follows.

Rearranging Eq. 4 to include DC loss and substitute variables with known variables (i.e., P_{loss} and R_{loss}), yields:

$$\eta_{sys} = \frac{P_{pa} - P_{in} - P_{loss}}{P_{dc} + P_r}$$

$$\eta_{sys} = \frac{P_{pa} - P_{in} - P_{loss}}{V_{dc} I_{dc} + I_{dc}^2 R} \qquad (16)$$

where:

P_{dc} = the DC power delivered to the PA,

P_{loss} = loss,

P_{pa} = generated PA output power,

P_{in} = input power, and

P_r = the dissipated power due to ohmic loss denoted as R.

And, finally substituting the equation with power terms in dBm, Eq. 17 describes the relationship between system efficiency and loss. As seen from the surface plot in **Fig. 5**, small losses quickly degrade system efficiency.

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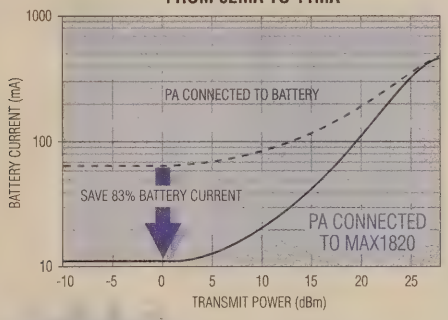
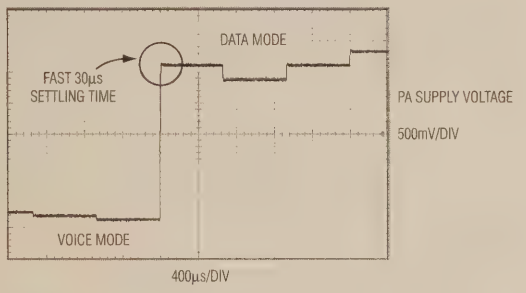


OUTPUT OF HIGH-SPEED DC-DC TRACKS PA POWER ENVELOPE WITHIN 30µs



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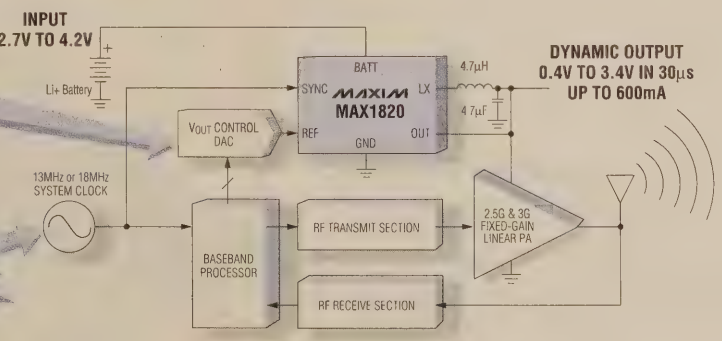

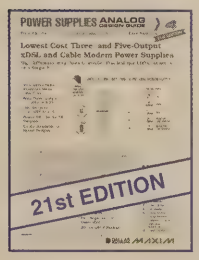

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$$\eta_{sys} = \frac{1\times 10^{-3}\left[10^{(P_{padb}-P_{lossdb})\times 10^{-1}} - 10^{P_{indb}\times 10^{-1}}\right]}{I_{dc}\left(V_{dc} + I_{dc}\times R\right)} \quad (17)$$

There are ways to either eliminate or minimize loss since loss often depends on layout and choice of architecture. It is not uncommon to have 200-mΩ ohmic loss and to have a 1.5-dB insertion loss in a mobile phone. Using these numbers reduces η_{sys} from 54 to 35 percent. This emphasizes the need to minimize RF loss.

There may be some disagreement with Eq. 17. Since there is no insertion loss, the output power of the PA would have to be backed off to get the same output power when loss is present. For example, if the insertion loss is 1.5 dB and the desired output power at the antenna is +33 dBm, the required PA output power would have to be 1.5 dB higher or +34.5 dBm. If this PA were being used in a theoretical system without loss, the efficiency would be less than the maximum efficiency at P_{out} = +34.5 dBm and, therefore, Eq. 17 would not hold true in practice. A more accurate way to describe system efficiency is to keep P_{out} at the antenna constant, while including the effect on PAE versus PA output power.

All amplifiers adhere to the same physical rules regardless of technology. For gallium-arsenide (GaAs) heterojunction bipolar transistors (HBTs), the saturation voltage (V_{sat}) depends on collector current I_{cc} primarily due to emitter resistance. Higher V_{sat} yields higher DC loss in the transistor and, thereby, generates less RF power, which results in lower PAE. This implies that lower output power is preferred due to the transformation ratio in the matching network diminishing as a function of P_{out}. As a result, the higher load line decreases the loss in the matching network. The relationship between load line, number of cells in the output stage, and PAE is beyond the scope of this article, but a linear relationship of 3 percent per decibel extra output power is realistic and will be used in the modified equation based on Eq. 4. The subtracted efficiency versus extra output power uses the linear relationship P_{lossdB}/α, where $\alpha = 30$. This relationship is a reasonable approximation for $0.48 \geq \eta \geq 0.62$.

Since P_{out}, or the delivered power to the antenna, is held constant, $P_{PA} - P_{loss}$ is equal to P_{out}. PAE is defined here as $\eta - P_{lossdB}/\alpha$ and I_{dc} is described as a function of α, PA output power, and insertion loss resulting in:

$$\eta - \frac{P_{lossdb}}{\alpha} = \frac{P_{outrf} + P_{loss} - P_{in}}{Pdc} \quad (18)$$

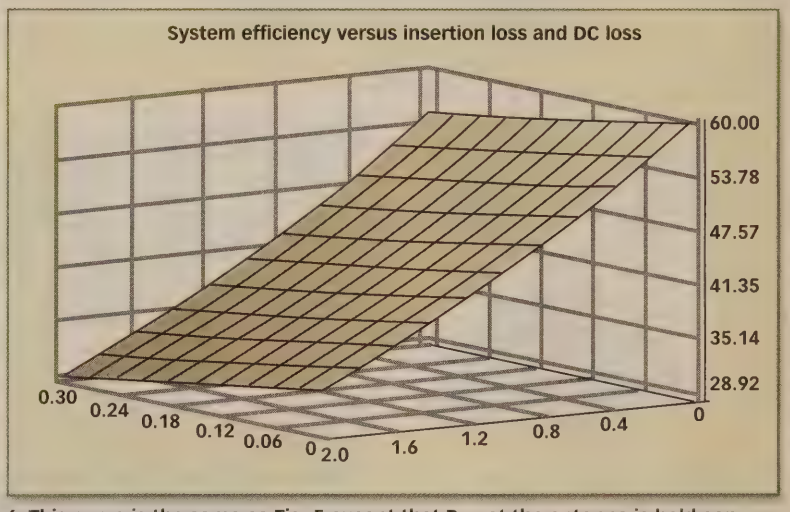
P_{outrf} is held constant at all times and a particular P_{DC} therefore exists as a function of P_{lossdb} and P_{loss}. If η is 0.6, and P_{lossdb} = 0, a particular P_{dc} can be calculated knowing the output power and P_{in}. Equation 18 is rearranged and isolated for I_{dc} and written in db and dBm terms.

(See Eq. 19 below.)

I_{dc} is then substituted into Eq. 7 and the total η_{sys} as a function of α, R, and P_{loss} becomes:

(See Eq. 20 below.)

Even though **Fig. 6** shows how the system efficiency



6. This curve is the same as Fig. 5 except that P_{out} at the antenna is held constant. The axes representations are the same as in Fig. 5

$$I_{dc} = \frac{P_{outrf} + P_{loss} - P_{in}}{V_{dc}\left(\eta - \frac{P_{lossdb}}{\alpha}\right)} \qquad I_{dc} = \frac{1\times 10^{-3}10^{(P_{outdb}+P_{lossdb})10^{-1}} - 1\times 10^{-3}\times 10^{P_{indb}\times 10^{-1}}}{V_{dc}\left(\eta - \frac{P_{lossdb}}{\alpha}\right)} \quad (19)$$

$$\eta_{sys} = \frac{1\times 10^{-3}\left[10^{(P_{outdb})10^{-1}} - 10^{P_{indb}10^{-1}}\right]}{\frac{1\times 10^{-3}10^{(P_{outdb}+P_{lossdb})10^{-1}} - 1\times 10^{-3}10^{P_{indb}10^{-1}}}{V_{dc}\left(\eta - \frac{P_{lossdb}}{\alpha}\right)}\left[V_{dc} + \frac{1\times 10^{-3}10^{(P_{outdb}+P_{lossdb})\times 10^{-1}} - 1\times 10^{-3}10^{P_{indb}10^{-1}}}{V_{dc}\left(\eta - \frac{P_{lossdb}}{\alpha}\right)}R\right]} \quad (20)$$

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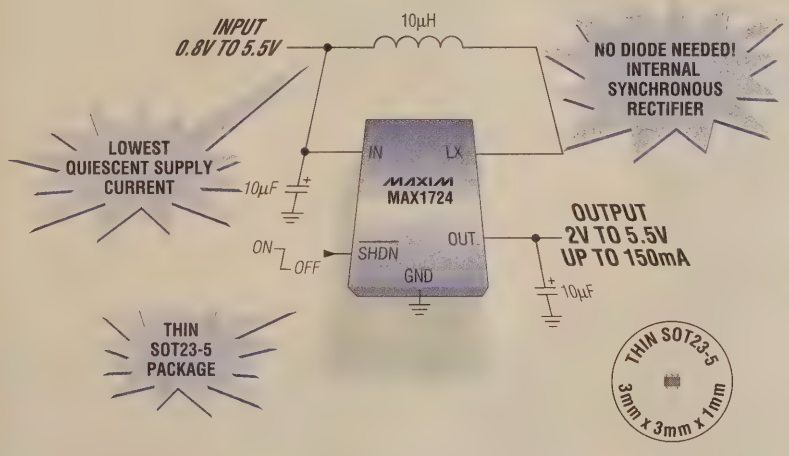

PART	OUTPUT* (V)	$\overline{SHDN}$	LOW-EMI CIRCUITRY
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MAX1723EZK	Adjustable	Yes	No
MAX1724EZK27	Fixed 2.7	Yes	Yes
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MAX1724EZK33	Fixed 3.3	Yes	Yes
MAX1724EZK50	Fixed 5.0	Yes	Yes

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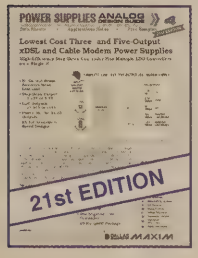

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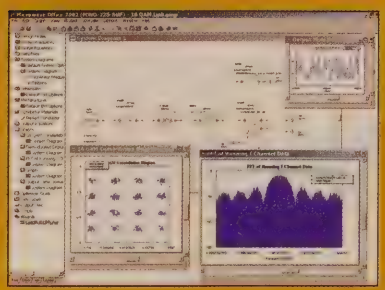

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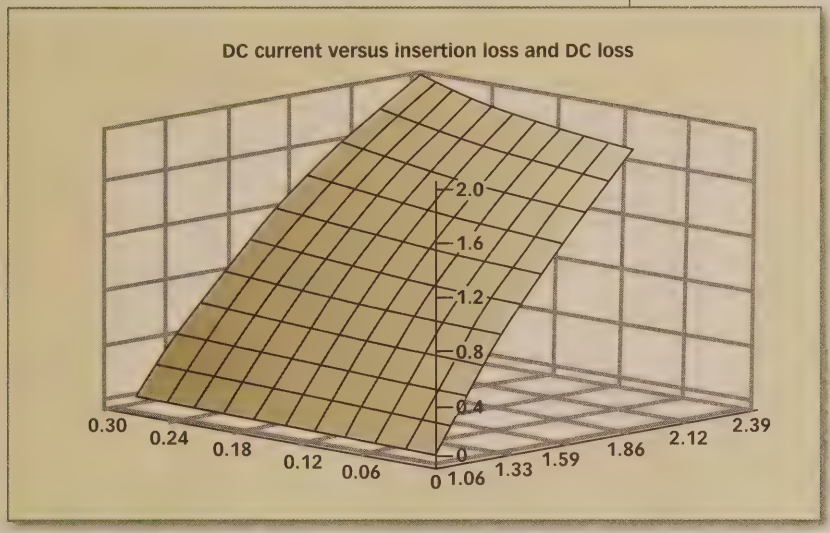
degrades as a function of insertion loss and supply-side loss, it does not indicate how it impacts talk time.

If a PA presents +33.5-dBm output power at the antenna and no loss is present, it is possible to achieve 60-percent efficiency, as shown in Fig. 6. If there is 1.5-dB insertion and 200-mΩ DC loss, this number decreases from 60 to 35.6 percent.



7. DC current versus insertion loss and DC loss is shown in these curves. The x-axis is ohmic loss from 0 to 0.3 Ω, the y-axis is insertion loss from 0 to dB, and the z-axis is DC current.

Current draw is used when calculating talk time, but why use the word "efficiency" when it seems to have no relevance? Efficiency is used as a number that helps designers know how optimized their designs actually are. In some cases, there is only one solution or architecture that applies to a particular system. In those cases, it helps the designer evaluate system performance. But for mobile-telephone designers, talk time is a critical measure. It is also easily comprehended by mobile subscribers and helps them compare the talk time of competing manufacturers' phones. Even though talk-time calculations are often very comprehensive and require knowledge of idle currents, current draw, and power-on sequences, it is still possible to calculate talk time with decent accuracy. As previously mentioned,

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a system's or component's current draw is needed to define talk time.

If P_{DC} is substituted in the denominator of Eq. 18 with $(V_{bat} - R \times I_{dc})\, I_{dc}$ the result is:

$$(V_{bat} - RI_{dc})I_{dc} = \frac{P_{outrf} + P_{loss} - P_{in}}{\eta - \frac{P_{lossdb}}{\alpha}} \qquad (21)$$

This is a second-order equation. Arranging in decreasing order results in:

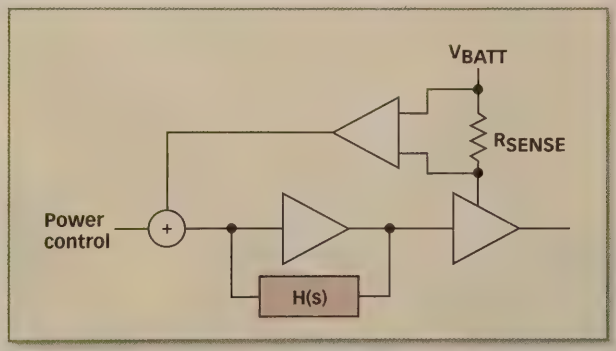
$$-RI_{dc}^{\,2} + V_{bat}I_{dc} - \frac{P_{outrf} + P_{loss} - P_{in}}{\eta - \frac{P_{lossdb}}{\alpha}} = 0 \qquad (22)$$

Equation 22, which describes I_{dc} as a function of DC loss and insertion loss, is plotted in **Fig. 7.** This figure shows the devastating effect of loss on current draw and that the impact on talk time could be very severe even with few losses. Looking at some typical numbers for a mobile telephone where P_{loss} = 1.8 dB and R = 200 mΩ yields I_{dc} = 2.02 A. If the DC and insertion losses are zero, then I_{dc} is 1.06 A. This means that approximately 1000 mA times the battery voltage is dissipated as unusable heat.

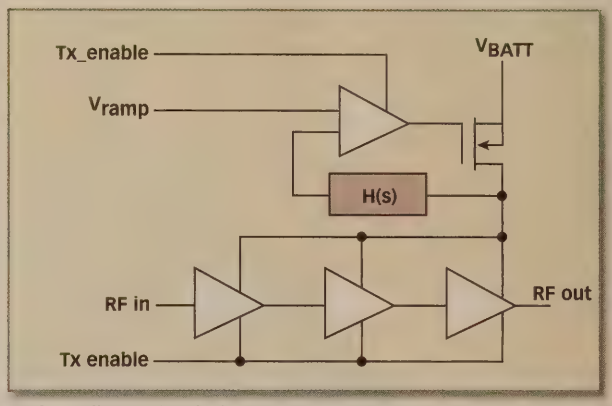
The argument can be made that the example is too optimistic since loss is inherent and can never be eliminated, but the next section will focus on power-control architectures and their impacts on power consumption.

Directional-Coupler Architecture

The mobile-phone designer has several choices when deciding which power-control architecture to use. Using a directional coupler and feeding it back to an error amplifier to sense forward incident power is probably the most common power-control method, and has been used for decades in various types of Txs. One advantage of the directional coupler is that the forward power is kept constant, while the dissipated power varies with a function of load impedance. The disadvantage of using a directional coupler, especially



8. Current-sense architecture is a technique to eliminate post PA losses that are associated with a directional coupler. Choosing the correct sense resistor is important to prevent losses and lower efficiency.



9. The collector (or drain) power-control method depends on the bipolar or FET's saturation voltage V_{sat}. The lower the saturation voltage is, the lower the loss is and the more efficient the supply is.

if it is printed on the printed-circuit board (PCB), is its physical size and possible electromagnetic (EM) radiation problems. Insertion loss in the directional coupler mainly depends on layout and material properties. A typical number for insertion loss including mismatch loss at 900 MHz is 0.3 to 0.5 dB (Table 1) and at 1800 MHz is 0.4 to 0.6 dB (Table 2). Lowpass filters are used to filter the harmonics generated in the PA. Depending on the amplitude of these harmonics, a third-to-fifth-order filter is typically required, which means an additional 0.5-to-0.7-dB insertion loss. In addition, an antenna switch is also required to switch between receiver (Rx) and Tx mode with a typical loss of 1 dB. Adding all of the losses together provides 1.5 to 2.0 dB. Eliminating the directional coupler makes it possible to reduce the loss by shaving approximately 0.5 dB off the total loss, indicating that the power-control architecture investigated is less optimal.

Although most power-control-loop designs implement a forward-power detect circuit, some power-control systems use a supply-current detect circuit. The principle for this circuit is to control current. By regulating current, the power is also regulated. This approach is appealing to designers due to the elimination of the post PA loss associated with a directional coupler and, hopefully, a reduction in current. However, the selection of the sensing resistor is dictated by the part with the lowest current, which is the DCS amplifier, and implies a sensing resistor of 100 mΩ **(Fig. 8).** This is done to maintain sufficient power-control resolution at low power levels. In GSM mode, the voltage drop across 100 mΩ is approximately +0.2 VDC and incrementally reduces the efficiency by approximately 4 percent.

A collector or drain voltage-control architecture takes advantage of the linear relationship between voltage and output power and uses a field-effect transistor (FET) to regulate the voltage applied to either the collector or drain on the PA. Using a proper design, a directional coupler is not needed to keep the output power constant. Therefore, by definition, only DC

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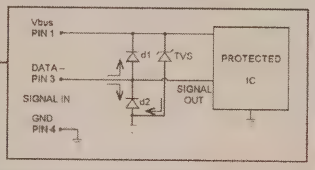

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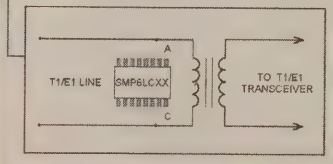
APPLICATIONS

- T1/E1 Protection
- Data line Protection



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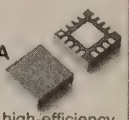

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FEATURES

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- 27 dBm Linear Power Output
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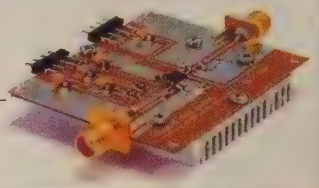
- 40% Linear Efficiency
- 70mA Idle Current
- 3mm 16 Pin MLP or Flip Chip

APPLICATIONS

- 1920-1980 MHz W-CDMA Handsets
- 1850-1910 MHz CMDA2000 Handsets
- Spread Spectrum Systems



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APPLICATIONS

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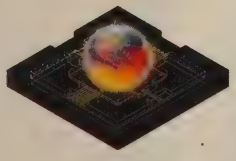
loss exists and is determined by the on-resistance in the FET. With a proper process, chip layout, and device properties, the on resistance can be less than 15 mΩ. This approach has the lowest effective loss of the three power-control architectures presented in this article (**Fig. 9**). Now that the optimum power-control architecture has been identified, the question still remains whether something can be done to the PA to reduce its losses.

A parameter, which has been used among PA chip designers for decades to describe PA performance, has also been used frequently in this article. This parameter is PAE. In general, chip designers tweak and modify the PA until the last tenths of a percent improvement is obtained. But does this actually yield the optimum solution or performance? In terms of efficiency, it is yes. But what about other parameters, such as harmonics? Harmonics that are generated in the PA often force the user to add a lowpass filter with approximately 0.3 to 0.5 dB of insertion loss. It would be best if this loss could be eliminated completely. For a moment, assume that complying with harmonics can be achieved using a simple matching network. Although this network may degrade efficiency, the question is by how much? The analysis of matching networks that minimize harmonics while maintaining good efficiency is beyond the scope of this article. However, it seems that harmonics can be met, degrading efficiency by 2 percent while the total system efficiency actually increases 3 percent. These numbers clearly emphasize that optimizing the PA for maximum PAE often turns out to have a negative impact on total system efficiency.

As PAs become more integrated and more loss is added, designers have the benefit of controlling these losses, including mismatch losses. It also means that from a PA perspective, PAE is no longer valid simply because PAE is a measure of the amplifier performance only, excluding all other circuits. Furthermore, when comparing different PA solutions, one has to consider what is being compared. If a PA module with integrated power control offers 53-percent efficiency and another module without power control offers 55-percent efficiency, which of the two modules then has the best system efficiency?

If a PA without integrated power control is placed in an external power-control circuit, what must the required PAE be to obtain similar system efficiency without power control? It is necessary to introduce a new term that describes efficiency regardless of system architectures. This term is "effective efficiency" or η_{eff}. Even though the definition is confined to using power con-

Continued on page 125



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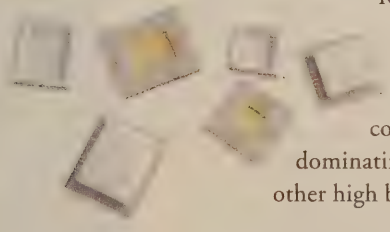

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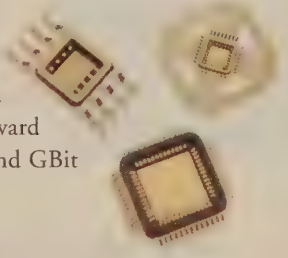

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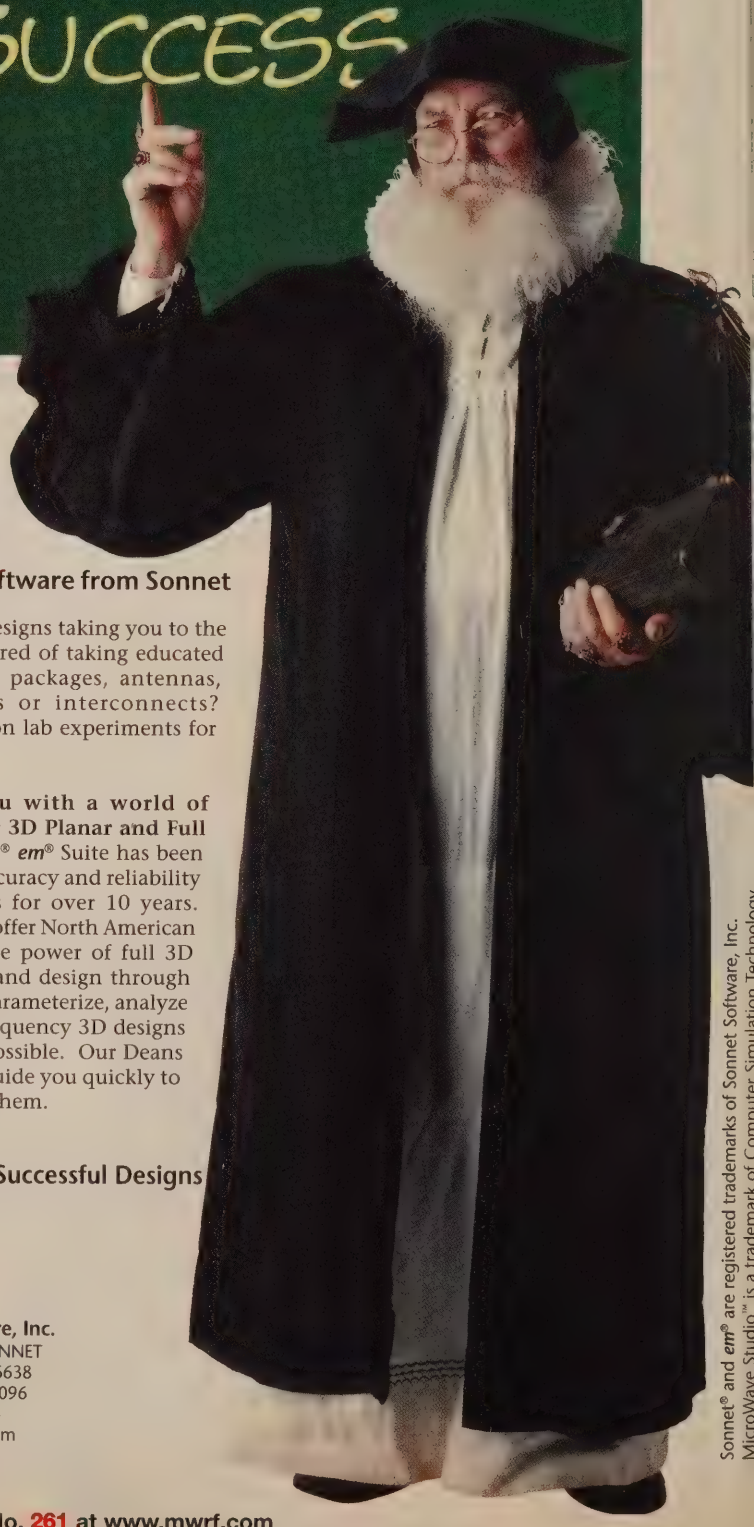

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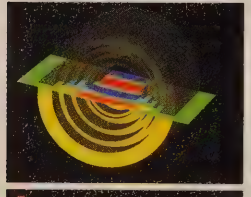

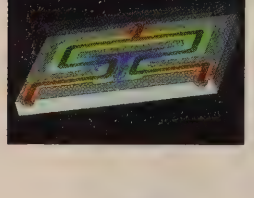

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Understanding Regulations For Short-Range Radios

The choice of operating frequencies for a short-range radio system is usually based on controlling where harmonically generated signals fall and avoiding possible sources of interference.

FARRON L. DACUS
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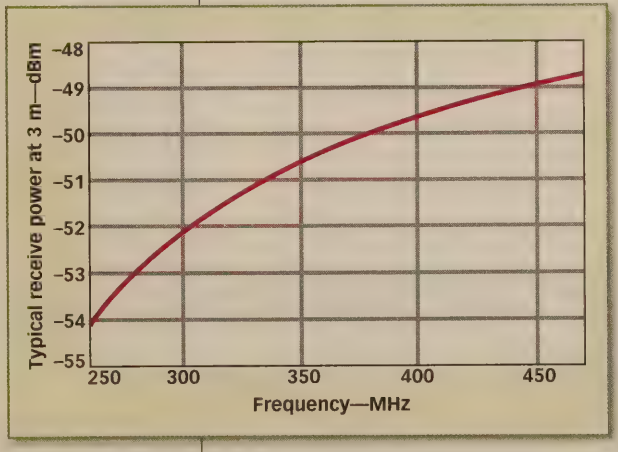
Short-range radios provide the means for function control and limited data access between electronic systems. But these radios must perform in accordance with applicable regulatory agencies, such as the Federal Communications Commission (FCC) in the US and the European Telecommunications Standard Institute (ETSI). Part 2 of this four-part article series on short-range radios will review various global radio-transmission regulations and their implications for the design of low-cost short-range radios.

Classical short-range radios for control and security applications are typically in the 300-to-500-MHz range, with transmit-power levels from −30 to +10 dBm. These radios are generally certified, a form of approval where the manufacturer has the product tested in an approved laboratory that will confirm that the product meets specifications, and assists in submitting a report to regulatory agencies such as the FCC. The regulatory agency will then grant approval for production, and the end customer will never have to deal with any licensing issues. European and US rules require that the antenna of the certified equipment be integrated or use a customized connector to prevent users from substituting other antenna types.

The FCC rules officially govern operation only in the US, but also are adopted in varying degrees by many other nations in the Americas and the Pacific rim. There are four specific FCC rules



1. This plot shows the change in link budget (received power) with frequency when following FCC 15.231 rules, with an approximate 5-dB improvement in link budget over frequency.

of high interest to designers of control and security class short-range radio systems, with interrelationships between the four that must be understood. These four sections are 15.209, 15.231, 15.205, and 15.35. The rules for industrial-scientific-medical (ISM) band radios are provided in 15.249 for 1-mW (0-dBm)-class narrowband systems, and 15.247 for spread-spectrum systems up to 1 W (+30 dBm).

$$E_{SS}(f) = 0.041667(f - 260) + 3.75(\textit{control operations}) \quad (8)$$

The so-called "general" rule, 15.209, restricts the RF energy that electronic equipment may parasitically emit. The specific level of emissions is 200 μV/m at 3-m test range below 960 MHz, and 500 μV/m above. These field strengths are equivalent to approximately −49- and −41-dBm effective radiated power (ERP), respectively.

Rule 15.231 is the major authorization for control and security-class equipment. For control applications, the permitted peak transmit level varies linearly from 260 to 470 MHz as provided by Eq. 8,

where:

E_{SS} = the permitted steady-state field strength [in root mean square (RMS) mV/m at 3 m for control applications], and

f = frequency (in megahertz).

For frequencies above 470 MHz, the permitted power is equal to that at 470 MHz. Equation 6 can be used to show that the transmitted ERP range is −24 to −13 dBm. These increasing FCC permitted power levels actually exceed the effects of decreasing antenna aperture with increasing frequency **(Fig. 1)**. The band is approximately 5 dB better at the high end than at the low end, a fact that is apparently not well-known, but one whose positive effect may be reduced by the interference of television second harmonics. The permitted harmonic levels are 20 dB below these levels, except where they fall in restricted bands (below). Note that television-broadcast channel 13 ranges from 210 to 216 MHz, and then allocation skips to channel 14 at 470 to 476 MHz, so direct-television interference is not a problem throughout this range (though second harmonics can be a problem). The permitted RMS steady-state field strength for periodic operation (such as status reporting) not limited to control is approximately 8 dB lower. In addition, it must meet timing restrictions (except in emergencies) by having an off time of at least 30 times the transmit time, with transmissions that

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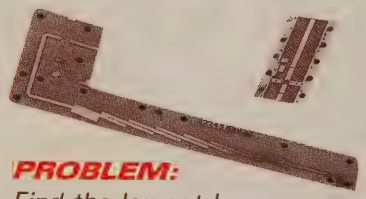
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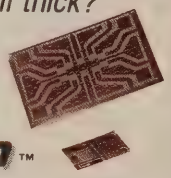
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UMX-270-D16	2160-2350	0.5-4.5	80	1.1:1	+5, ±2	-20	-106	0.7	2	5
UMX-315-D16	2175-2175	0.5-4.5	7	1.05:1	+7, ±2	-20	-120	0.5	2	6
UMX-333-D16	2650-2950	1-14	30	1.05:1	+5, ±2	-20	-104	1.0	3	6
UMX-375-D16	2850-2850	0.5-4.5	7	1.05:1	+7, ±2	-20	-118	0.8	2	6
UMX-331-D16	3125-3275	0.5-4.5	50	1.05:1	+5, ±2	-20	-104	1.0	3	6

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do not occur more often than once per every 10 s. (Under FCC 15.205, 260 to 285 MHz falls into a restricted band and may not actually be used.)

Section 15.205 of the FCC regulations documents the "restricted bands" where only spurious emissions are permitted, and where those must meet the general levels of 15.209. Above 1000 MHz, averaging as described in FCC 15.35 may be used. From these restricted bands, a list of desirable frequencies can be formed **(Table 1)**.

From a link-budget point of view, the most desirable segment is 410 to 470 MHz. This section may be used at the maximum allowed power if approximately 28 dB of third-harmonic rejection is attained. Frequencies above 432 MHz also have the virtue of avoiding television harmonics. If a simple unfiltered loop antenna, required by cost considerations, does not attain this rejection, then an option is to use the reduced segment from 413.33 to 433.33 MHz. This segment is only 1 dB below the maximum link budget, and places no harmonics below the sixth harmonic in restricted bands. However, the 420-to-450-MHz range is an amateur-radio band. There is some land-mobile-radio operation from 421 to 430 MHz, and 418 MHz is a popular surface-acoustic-wave (SAW)-based frequency that some users might wish to avoid when possible. Based on this potential interference, the segments from 413.33 to 417.90 and 418.1 to 420.0 MHz would be superior. These segments do, however, fall into the second harmonic of television channel 12. The 307-to-320-MHz range would be preferred as a band that avoids direct interference and television harmonics, at the cost of

Table 1: Primary restricted frequencies under FCC 15.205

RESTRICTED FREQUENCY RANGE (MHZ)	IMPACT
240 to 285	No fundamental usage here
322.0 to 335.4	No fundamental usage here
399.9 to 410.0	No fundamental usage here
608 to 614	Medical telemetry band, second harmonics from 304 to 307 MHz must meet the general requirement in this segment
960 to 1240	Note that third harmonic of 320.00 to 413.33 MHz must meet general
1300 to 1427	433.33 to 470.0 MHz must meet general with third harmonic
1435.0 to 1626.5	287.0 to 325.3 MHz must meet general with fifth harmonic

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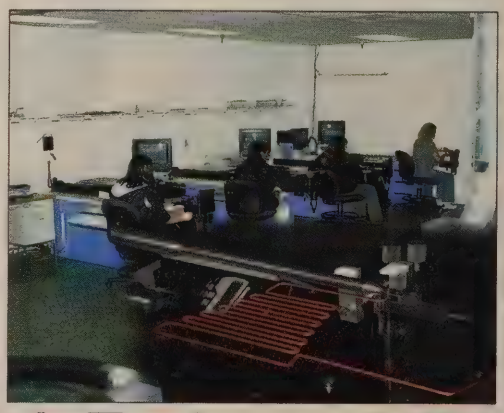

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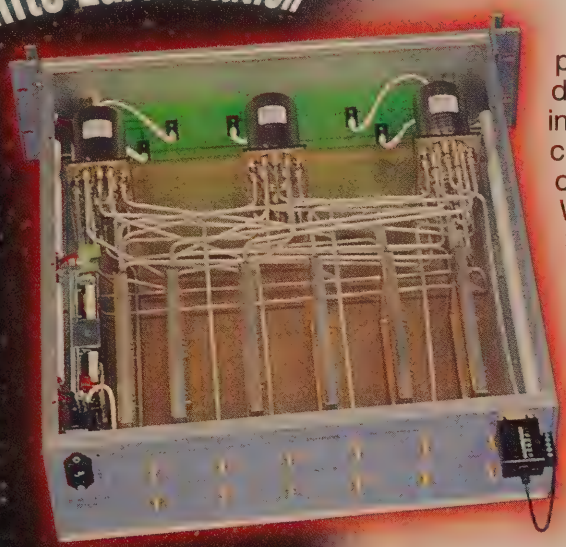

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requiring the fourth harmonic of 307 to 310 MHz to meet the general level specification (500 μV, −41-dBm ERP) and suffering approximately a 3-dB link-budget degradation compared to the high end of the band. The subrange from 310 to 320 MHz is clear of television harmonics and its own harmonics dodge restricted bands through the fourth harmonic.

FCC section 15.35 covers the permitted increase of peak transmit-power levels when averaged with off times that use amplitude-shift-keying (ASK) modulation. The use of this averaging is popular in the US due to an apparent accidental misinterpretation of the rules that have since become accepted as standard practice. The original intent of the rules was probably to support averaging for maintaining constant energy per bit. However, due to the FCC standard practice of using electric-field strength instead of ERP, the rule was, in its early usage, interpreted to allow for the maintenance of average field strength, up to 10 times the maximum steady-state field strength. Since power is increasing as the square of field strength, this permitted peak power to increase up to 100 times, and average power to increase up to 10 times. The decrease in transmit time results in an increase in signal bandwidth that is exactly proportional to the increase in transmit power. There is no increase in signal-to-noise ratio (SNR) if receiver (Rx) bandwidth tracks. However, there are two exceptions to this, which will be discussed later.

The precise mathematics of ASK averaging may be derived in short order, using the following definitions:

D_C = the total digital duty cycle (FCC uses the 100-ms segment in the protocol with highest duty cycle). Note that if the bit density is 50-percent digital "ones" and if transmission is not made during digital zeroes, then the duty cycle is down to 0.5 even before any bit shortening of digital ones is applied.

D_{C1} = the duty cycle of each individual "one" counting bit shortening. Note that if the length of a one were cut in half (such as standard Manchester), then D_{C1} = 0.5 and with 50-percent digital ones, D_C = 0.25.

D_{C0} = the duty cycle of each individual digital zero. For simple ASK, this could be 0; for standard Manchester it is also 0.5; for a nonstandard Manchester, it could vary from 0 to 1 depending upon averaging desired (which also applies to D_{C1}).

D_0 = the logical duty cycle of zeroes in bit stream, typically 0.5.

D_1 = the logical duty cycle of ones in bit stream, typically 0.5.

Now, for the total duty cycle (fraction of time carrier is transmitted):

$$D_C = D_O D_{C0} + D_I D_{C1} \qquad (9)$$

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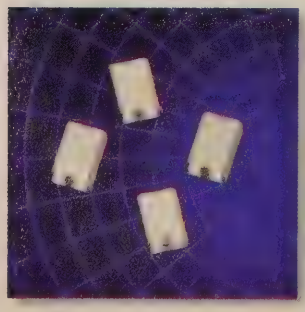
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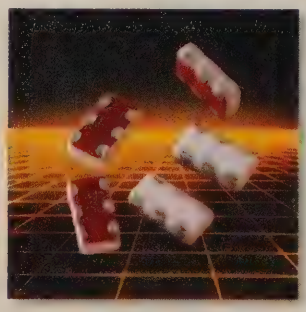
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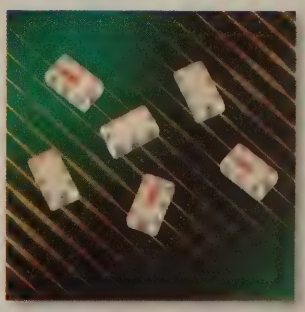
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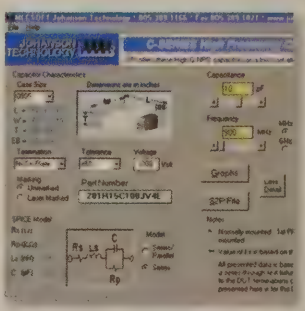
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Eq. 10 can be shown by defining:

E_{SS} = the field-strength steady state, which equals the permitted RMS field strength at a particular frequency when not averaging, and

E_{pa} = the field-strength peak averaging, the permitted peak field strength at a particular duty cycle and frequency when averaging. According to FCC convention, this "peak" is not the true RF peak. It is the RMS carrier strength in volts per meter at the peak of the envelope:

$$E_{pa} = E_{SS} / D_C \qquad (10)$$

which shows the maximum permitted peak field strength under the rules, to a limit of 10 times the steady state allowed (E_{SS}).

Eq. 9 can be substituted into Eq. 10 and solved in Eq. 11 (p. 86) for the necessary duty cycle within the bits given permitted (regulatory) and attainable (hardware-limited) field-strength levels. This can be shown in power terms as Eq. 12 (p. 86).

The averaging effect may be used to reduce harmonic-attenuation requirements above 1000 MHz, where "in the restricted bands," a level of 500 μV/m is generally required. The 500-μV/m level corresponds to 7.5×10^{-8} W ERP or −41.2-dBm ERP.

The effect of averaging strongly influences the choice between ASK and frequency shift keying (FSK) in FCC follower countries, which may be quantified as follows. For the case of nonreturn-to-zero (NRZ) ASK data with possible pulse shortening, one may write for the average ASK power, P_{ASK}:

$$P_{ASK} = 0.5 D_{C1} P_{pa} \qquad (13)$$

In Eq. 13, $D_{C1} = 1$ if no pulse narrowing is not used. If the Rx bandwidth is the reciprocal of the symbol time with pulse narrowing, T_{NS}, and without pulse narrowing, T_S, and taking into account the 50-percent duty cycle of simple ASK, then it may be shown that the ratio of the SNRs for ASK transmit power, P_{pa}, between 0 and 6 dB over the steady-state limit, P_{SS}, and FSK transmit power at P_{SS} is shown in Eq. 14 (on p. 86).

This advantage scales from 0 to 3 dB as P_{pa} advances from 0 to 6 dB over P_{SS}. Substituting Eq. 12 for Eq. 14 will show that for P_{pa} greater than 6 dB over P_{SS}, the advantage for ASK tops out at 3 dB, so as long as Rx bandwidth is at the reciprocal of symbol time. However, low-cost Rxs for this class of equipment are often unable to match their noise bandwidth to be only the reciprocal of symbol time so, in practice, averaging is often a significant improvement.

It may be apparent that US rules for control and security operation are not particularly easy to follow, being spread over four sections, lacking in some definitions, and not written in a tutorial fashion. However, they are pleasant reading compared to the European rules, which are spread out over many documents, often seem to lack references to essential supporting data and, in general, seem to feature considerable disregard for reader convenience. These rules have been

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$$D_0 D_{C0} + D_1 D_{C1} = E_{SS} / E_{pa} = (allowed\ electric\text{–}field\ strength) / (attainable\ electric\text{–}field\ strength) \quad (11)$$

$$D_0 D_{C0} + D_1 D_{C1} = \left(P_{SS} / P_{pa}\right)^{0.5} = (allowed\ steady\text{–}state\ power / allowed\ peak\ power)^{0.5} \quad (12)$$

$$SNR_{ASK} / SNR_{FSK} = \left(0.5 D_{C1^2}\right)\left(P_{pa} / P_{SS}\right) \quad (14)$$

reviewed here in as practical a way as possible, although there may be mistakes made in these interpretations of the European regulations. Fortunately, there is a fairly high degree of standardization on bands and transmit-power levels throughout Europe, with most disagreements coming in the permitted modes and transmit duty cycles. This is achieved under the regional authority of the European Conference of Postal and Telecommunications Administrations (CEPT), which has 43 member nations. ETSI develops technical standards for CEPT countries.

A significant philosophical difference in the European rules are provisions that go beyond preventing interference to other systems attempting to guarantee acceptable system performance. Most electronic equipment sold in the European Union must comply with EMC Directive 89/336/EEC and be labeled with the CE mark, in conformance with this policy. After April 8, 2000, compliance with these requirements could be self-certified by certain procedures (see www.ero.dk). Another important document governing required performance its ETS 300 683, "Electromagnetic Compatibility (EMC) Standard for Short Range Devices Operating Between 9 kHz and 25 GHz." This document's performance requirements are centered on interference immunity from outside EM fields and disturbances on power-supply and control inputs. The document refers to other documents for many measurement details. Designers of finished radio equipment to be marketed in Europe must review this set of documents in some detail.

For control and security applications the most fundamental document is CEPT ERC Recommendation 70-03 (which can be downloaded at www.ero.dk). Further details on test methodology to confirm compliance are provided in ETSI EN 300 220-1 (which can be downloaded from ETSI's site at www.etsi.org). EMC compliance is described in ETSI ETS 300 683. For ISM-band operation, which generally includes higher-end applications such as Bluetooth and wireless local-area networks (WLANs), ETSI 300 328 is, in general, the applicable document. Europe does not have the 902-to-928-MHz ISM band, but does use the same 2400-to-2483.5-MHz ISM band that is authorized by the FCC, though at a reduced spread-spectrum power level of 100 mW (the US is up to 1 W). Note that up to 10-mW ERP narrowband is authorized in the European 2400-MHz band, though this is not mentioned in ETSI 300 328 with the other 2400-MHz rules. Instead, this is authorized in CEPT recommendation 70-03 under Annex 1 (see **Table 2** and references).

For classic control and security applications that are similar to FCC 15.231, the European rules support the use of 433.05 to 434.79 MHz under the "non-specific short-range device" rules of Annex 1 under ERC 70-03E. This band segment just misses the second harmonic of television channel 13, a likely reason for its selection. The approval process is similar to US certification where individual licenses are not required. The basic rules here are up to 10-mW ERP, at less than 10-percent duty cycle, with some countries having differing duty-cycle limits. This generous transmit power has the capability to provide an excellent short-range link. Though applications are not specifically limited, the 10-percent duty-cycle limit inherently restricts applications to control, intermittent status reporting, and low-end data acquisition (DAQ). There is no designated frequency-accuracy specification provided, but since the band is not channelized, the general wide band ±100-PPM requirement of Table 7 in ETSI 300 220-1 (p. 24 of the regulation) should apply. This would appear to be a difficult requirement for untuned low-cost SAW-based devices to meet, but the apparent practical interpretation of this rule is that it covers temperature and power-supply drift, allowing tuned SAW-based devices to meet it with the assumption that set on error is negligible. The harmonics are not specifically called out, but the general spurious limits of 250 nW below 1000 MHz and 1 μW above 1000 MHz provided in Table 13 of ETSI 300 220-1 (on p. 34 of that regulation) should apply. Note that at full power of 10 mW, this is −46 dBc for the second harmonic and −40 dBc for higher harmonics. Since this limit is not set as strictly in dBc, for a typical ERP of −10 to −20 dBm attained with a printed-loop antenna, the limit to the second harmonic is closer to −26 to −16 dBc. Note that there are certain band segments where the operating-mode parasitics are limited to 4 nW, including 470 to 862

Table 2: These FCC 15.231 frequencies can be used to avoid carriers, second, and third harmonics in restricted bands below 1000 MHz

FREQUENCY (MHz)	COMMENT
285 to 304	Stop at 304 MHz to avoid placing second harmonic in medical band
307 to 320	Stop to avoid third harmonic from 960 to 1000 MHz. Note that the fourth harmonic of 307 to 310 MHz is restricted
335.4 to 399.9	Stop to avoid direct restricted
410 to 470	Power tops out at 470 MHz

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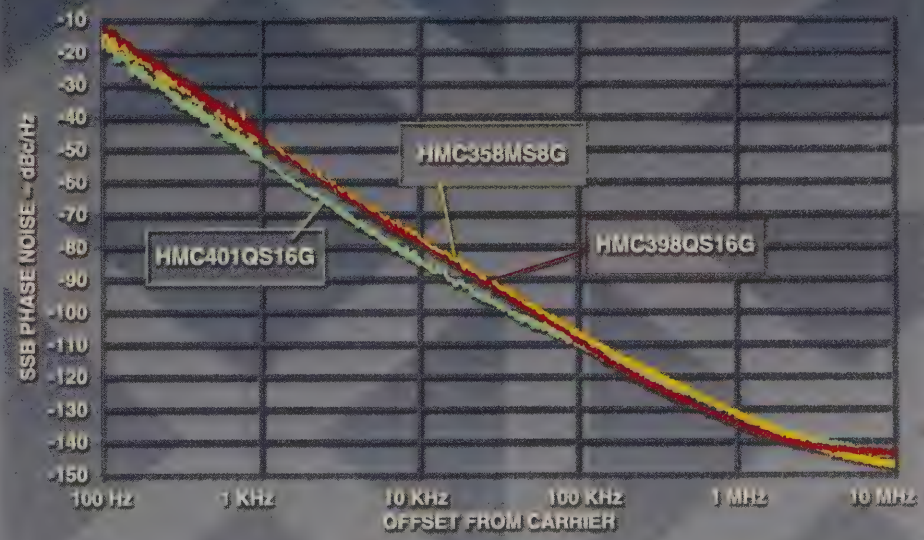

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MHz. At −10 dBm ERP, there is −44-dBc required spurious suppression in these band segments.

Although television-station frequency allocation skips over the US 260-to-470-MHz band, the frequencies are so powerful that their second harmonics are still a potential interference problem. Television stations 7 through 13 are permitted to have power levels to 316 kW, with harmonics for analog stations that are limited to −60 dBc or better (future digital TV is expected to feature much lower harmonics, reported as −110 dBc). **Table 3** shows television and FM channels and frequency allocations. The segment 285 to 348 MHz avoids television harmonics, and the segment 324 to 348 MHz avoids television and FM broadcast harmonics. However, the band from 322 to 335.4 MHz is directly restricted by FCC section 15.205, and there are restricted band-harmonic issues that make other segments less than optimum as well.

To gauge the potential problem, a standard handbook is referred to, such as ref. 3 for graphs of electric-field strength from broadcast stations. One approximation is 10 to 90 dBμV at the carrier frequency for distances from the broadcast transmitter (Tx) from 200 to 5 km per kilowatt of transmit power. Converting to watts for a 100-kW station with a Tx tower 100 m tall that are picked up at the −60-dBc second harmonic by a quarter-wave whip at 400 MHz shows a receive power of 1.7×10^{-21} to 1.6×10^{-13} W that is spread over the 12-MHz (at the second harmonic) television bandwidth.

Converting to dBm/Hz, the spectral density of this interfering source is approximately −225 dBm/Hz at 200 km up to −146 dBm/Hz at 5 km. The farther ranges are negligible as interference, but the shorter ranges definitely exceed the −174-dBm/Hz thermal noise floor that Rx sensitivities and link budgets are typically calculated against. With a typical inverse-forth power-propagation constant in this environment, interference could be expected up to approximately 20 to 40 km. This is conservative since the Rx antenna height is 9 m in the graph used. Actual power density may be 10 to 20 dB lower at the typical 1-to-2-m antenna height of short-range radio Rxs, reducing these interference ranges by a factor of two to four, but it does illustrate the potential seriousness of the issue. Some systems are likely to be suffering interference from this source that significantly reduces range and reliability, without designers being aware of the cause. Narrowband Rx systems can reduce this problem by using frequencies that are placed at the band edges of the television harmonics, such as 420 MHz, where energy density is approximately 20 dB lower. Fortunately, improved

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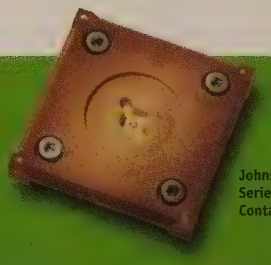

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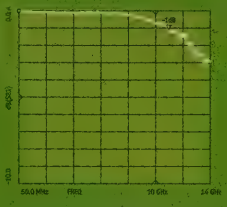

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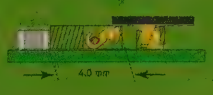

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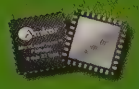

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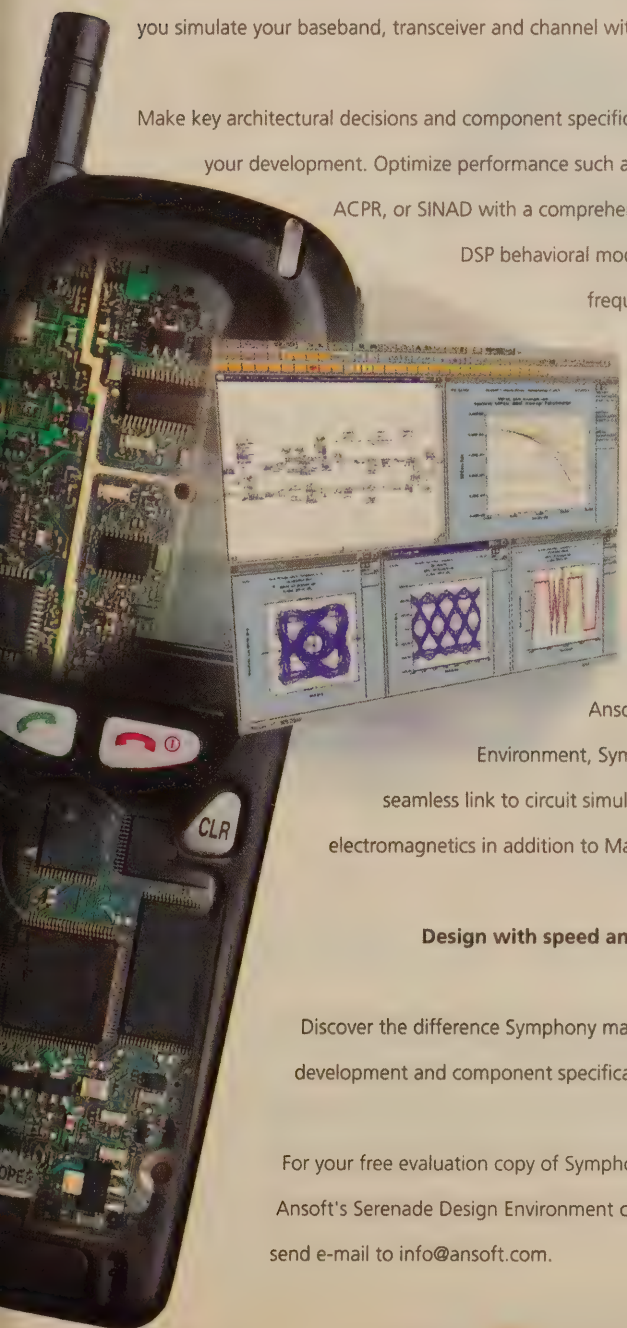

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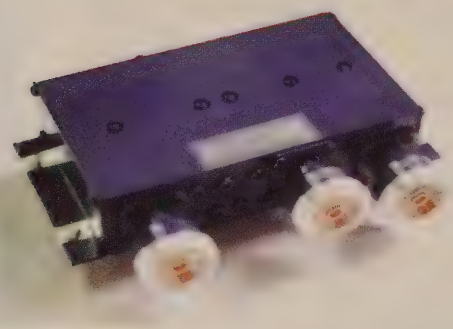

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filtering in newer digital television stations should reduce the incidence of this interference in the future.

The effects of regulations on frequency choices may be summarized as follows. The European choices are restricted to 433.05 to 434.79 MHz at 10-percent duty cycle (most countries) and a maximum power of +10 dBm among the lower ultra-high frequencies (UHF). If this is not adequate, the user may step up to 868 to 870 MHz with some segments, allowing up to +14 dBm at up to 100-percent duty cycle. The US choice of carrier frequency is a more complicated subject. The segments from 413.33 to 417.90 and 418.1 to 420.0 MHz are excellent for avoiding placing harmonics through the fifth harmonic in restricted bands that require more Tx filtering circuitry, while simultaneously dodging direct interference. They do suffer broadcast-television second-harmonic interference, the seriousness of which depends on location, frequency placement, and Rx bandwidth. The segment from 432 to 470 MHz is excellent for enabling near-maximum link budget and avoiding television and FM second and third harmonics, but does require up to −28-dBc harmonic rejection of its own third harmonic, which may be difficult to attain with an unfiltered loop antenna. The segment of 310 to 320 MHz is clear of television harmonics, and its own harmonics dodge restricted bands through the fourth harmonic. It does suffer from the third harmonic of FM broadcasting and its link budget is 3 to 5 dB inferior to the higher frequencies, but is suitable for low-cost applications that cannot pay for Tx filtering. In general, the 285-to-470-MHz band that is available in the US under FCC 15.231 is an under-used resource with many short-range-radio business opportunities.

By giving up segments of usable frequency to dodge "restricted bands" in the US, it is generally possible at typical transmit powers (0 dBm driving a 5-to-10-percent efficient loop antenna) to satisfy US and European rules with approximately 20 to 30 dB of harmonic suppression. For example, for US operation at steady-state output power in the 413.33-to-433.33-MHz band, only 20-dB suppression is needed through the fifth harmonic, and then 26 dB for the sixth harmonic. That same Tx, when reset to 433.92 MHz for the European market, needs at least 22 dB for all harmonics. If the European 0-dBm ERP Tx uses an efficient quarter-wave whip, then it needs a minimum of 36-dB second-harmonic suppression.

Despite the improved energy per bit that is possible with averaging in the US, the actual link quality remains little affected if Rx bandwidth tracks trans-

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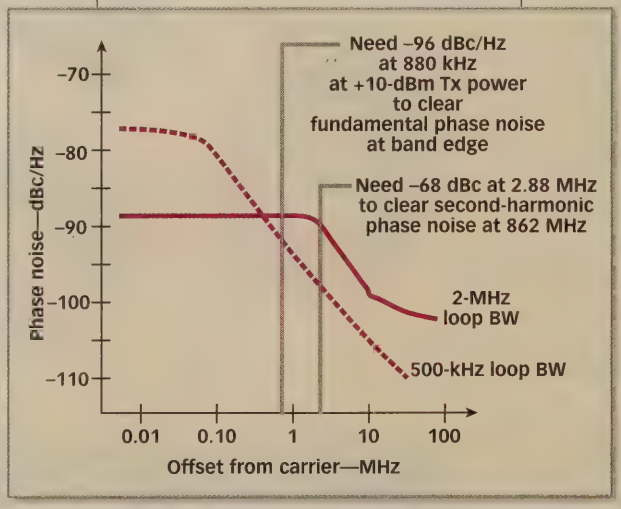
mitted spectral occupancy. An exception to this generalization is that ASK averaging without symbol shortening can yield an up to 6-dB improvement over nonaveraged ASK and up to 3-dB SNR improvement over FSK. In general, if narrowband Rxs can be provided in a particular application, better system design will result due to an equivalent link being maintained with more spectral efficiency (more channels and less interference) and with smaller Txs (smaller antennas and batteries). But if cost constraints force use of a wideband Rx, then averaging a shorter (wider-bandwidth) pulse of higher power will yield a significant link improvement for that nonideal case as compared to not averaging.

In high-mobility systems such as cellular networks, angle modulation [FSK or phase-shift keying (PSK)] is generally technically superior due to its improved multipath performance and greater interference immunity. Since Europe does not allow any ASK-friendly averaging effect, European FSK supports an inherently 3-dB improved link budget by the simple virtue of its 100-percent duty cycle. For low-mobility systems (most short-range radio applications) where Rx detectors are properly designed to provide a capture effect, FSK has little inherent link advantage. Therefore, in the US and other FCC countries, the choice between ASK and FSK, as far as link budget is concerned, depends on where the transmit power that a particular Tx, battery, and antenna may attain falls relative to the maximum permitted steady-state values. If the maximum attainable value is less than the steady-state limit, then FSK is superior since it will again show a 3-dB link-budget improvement. But if the maximum attainable power falls between the steady-state limit and 6 dB above the limit, ASK shows an advantage because the data duty cycle may be averaged without shortening pulses and widening Rx noise bandwidth. This net link improvement will be 3 dB over FSK if the ones bits may be transmitted at 6 dB above the steady-state limit. For transmit peak power that is more than 6 dB above the steady-state limit pulse shortening must be used, which widens Rx-noise bandwidth. This limits the advantage of ASK to at most 3 dB compared to FSK if the Rx bandwidth can be controlled to match transmitted spectral occupancy.

Table 3: Television and FM frequency assignments

CHANNEL	CARRIER (MHz)	SECOND HARMONIC (MHz)
6	82 to 88	164 to 176
7	174 to 180	348 to 360
8	180 to 186	360 to 372
9	186 to 192	372 to 384
10	192 to 198	384 to 396
11	198 to 204	396 to 408
12	204 to 210	408 to 420
13	210 to 216	420 to 432
14	470 to 476	940 to 952
FM stereo	88 to 108	Third harmonic = 264 to 324

The European rules, specifically ETSI EN 300-220-1 section 6.6, can also be inferred to set a phase-noise specification. Those regulations indicate that the spurious noise must meet the general European 250-nW spurious level (−36 dBm) using a 10-kHz spectrum-analyzer resolution bandwidth for carriers up to 25 MHz, using a 100-kHz analyzer-resolution bandwidth for carriers from 25 to 1000 MHz, and using a 1-MHz analyzer-resolution bandwidth for carriers above 1000 MHz. It would not be logical for this to be interpreted as a phase-noise requirement when the phase-noise offset frequency is still in an allocated band, such as 433.05 to 434.79 MHz (it is not a spurious product when it is in the permitted band), but it makes sense from the regulatory perspective to require integrated phase noise over these spectrum-analyzer bandwidths outside of the assigned bands to meet these spurious levels. Based on transmitted ERP, the inferred phase-noise requirement thus becomes Eq. 15 (p. 96)).



2. This plot compares the typical integrated Tx phase noise at two sample PLL bandwidths with the inferred phase-noise requirements for European (and US) short-range radio systems. At the full transmit power of +10 dBm, the phase noise fails by approximately 3 dB for the 500-kHz loop filter and about 8 dB for the 2-MHz loop filter.

At +10 dBm in the 434-MHz band, there is an implied out-of-band phase noise of −96 dBc/Hz. If a carrier is set at a nominal frequency of 433.92 MHz, then at 100-PPM maximum error it can get to within 880 kHz of the band edge, which is where this −96 dBc/Hz becomes applicable. The far-from-the-carrier transmitted phase noise (which may be helped by the frequency response of the antenna) must meet the 4-nW limit of the European "restricted" segments from 174 to 230 MHz and 470 to 862 MHz (Eq. 16).

(See Eq. 16 on p. 96.)

At +10 dBm and neglecting antenna bandwidth, this

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is a transmitted phase noise of −114 dBc/Hz.

It may also be inferred that the restricted band from 470 to 862 MHz, with its 4-nW spurious specification, sets phase noise and spurious limits for the second harmonic of the transmitted level, which could come in as low as 867.76 MHz with 100-PPM crystals. This implies that at 2.88 MHz offset from carrier, the transmitted phase noise must be at −68 dBc/Hz, and any discrete spurious must be 18 dB below the carrier if the second harmonic is at the 250-nW upper limit. These limits may usually be neglected since the carrier-frequency limits turn out to be the worst case.

This same rules section also sets the inferred synthesizer reference (phase-detector sampling rate) spurious specification shown in Eqs. 17 and 18 (see above).

The reference spurious component of −36 dBc for 0-dBm Txs is not difficult, but at +10 dBm and −46 dBc, care is required. The phase-noise specifications are not difficult for good-quality discrete designs, but may become more troublesome for low-quality-factor (Q) integrated designs, particularly at +10 -dBm ERP. This leads to a need for high loop bandwidths to suppress phase noise, which then tends to worsen synthesizer spurs. **Figure 2** shows phase-noise limits derived from these rules compared to typical performance of an integrated Tx at two different phase-locked-loop (PLL) bandwidths. The inferred mask, which is suitable for use in the US, is from a regulatory perspective. Acceptable performance for narrowband FSK in terms of phase-noise-limited SNR is an issue that shall be addressed in the next part of this article series. MRF

$$\phi_N(close\,in) = -\left[P_{ERP}\,(dBm) + 36 + 10\log(BW)\right] \quad (15)$$

$$\phi_N\,(far\ out) = -\left[P_{ERP}(dBm) + 54 + 10\log(BW)\right] \quad (16)$$

$$-\left[P_{erp}(dBm) + 36\right]\ for\ spurs\ in\ the\ 250\text{–}nW\ segments \quad (17)$$

$$-\left[P_{erp}(dBm) + 54\right]\ for\ spurs\ in\ the\ 4\text{–}nW\ segments \quad (18)$$

ACKNOWLEDGMENTS

The author wishes to express appreciation for the valuable reviews and inputs from other members of the Microchip RF Products Group, and from Didar Islam, David Saar of Saar Associates, Alan Bensky of Ateret Radio Engineering, Paul True of Linx Technologies, Frank Perkins of RF Monolithics, and Andreas Laute of Melexis.

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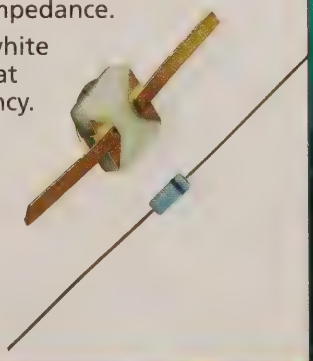

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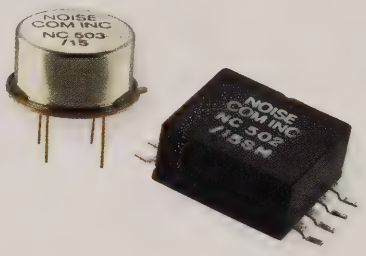

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NC502/15	0.2 MHz – 1 GHz	31 dB
NC503/15	0.2 MHz – 2 GHz	31 dB
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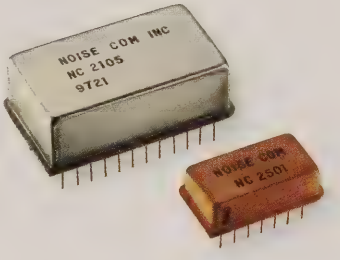

NC2000 SERIES
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MODEL	FREQUENCY RANGE	OUTPUT
NC2101	100 Hz – 20 kHz	0.15 Vrms
NC2105	500 Hz – 10 MHz	0.15 Vrms
NC2201	1 MHz – 100 MHz	+5 dBm
NC2601	1 MHz – 2 GHz	-5 dBm

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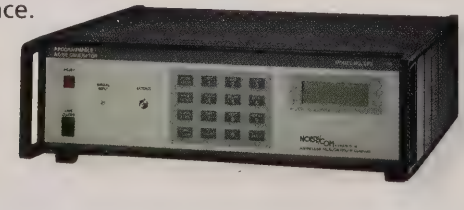

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MODEL	FREQUENCY RANGE	OUTPUT POWER
UFX7107	100 Hz – 100 MHz	+13 dBm
UFX7108	100 Hz – 500 MHz	+10 dBm
UFX7112	1 MHz – 2 GHz	+0 dBm
UFX7218	2 GHz – 18 GHz	-20 dBm
UFX7911	5 MHz – 1 GHz	+30 dBm

Use A Sampling Power Meter To Determine The Characteristics Of RF And Microwave Devices

The second part of this article discusses triggering, DAQ, and peak-mode measurements using a sampling power meter.

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Modern communications systems are placing increasingly greater demands on RF design and manufacturing engineers. Finished systems need to be tested to ensure compliance to exact international standards. Component parts are also increasingly tested using the same types of complex signals as their target application. Wireless-system designers now require full characterization of RF components and subsystems under dynamic signal conditions before they can qualify them for use in their designs. The old stalwart of RF measurements, the standard power meter, is creaking under the strain of these new requirements.

Part I discussed the digital storage oscilloscope (DSO), presenting its capabilities from the user's perspective as well as reviewing developments in DSO technology. Part II will present the capabilities of the sampling power meter from the user's perspective and review its technological developments. Sample measurements complete the article.

User's Perspective

Now with a full understanding of the basic architecture of a DSO, this discussion returns to the sampling power meter and contrasts its requirements.

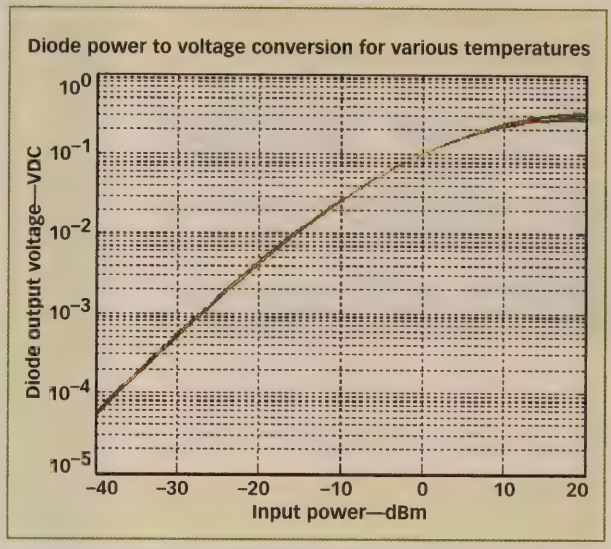
A sampling power meter digitally samples a signal that is derived from the envelope power of the RF input to the sensor. These data are then acquired, or made available for measurement, under the user's control. Several methods and controls allow the user to make different measurements. The essential point to identify at this time is that the sampling power meter is a measurement instrument. Its main purpose is to provide numeric certainty to an uncertain RF world, and great care is taken in these instruments to ensure measurement fidelity. This is in contrast to the DSO, where the emphasis is on the appearance of the waveform.

Compared with a DSO, a sampling power meters has the following distinctions:

- It is a measurement instrument.
- It is capable of nontriggered measurements.
- It is less display oriented.
- It requires complex signal conditioning for the sensor input.
- It requires high instantaneous dynamic range.

Making peak-mode measurements essentially requires the user of the sampling power meter to control the same parameters as the DSO user (i.e., display, trigger, and measurement).

Since the sampling power meter is a measurement instrument that is designed



7. This graph shows the power-to-voltage conversion of a typical diode.

primarily to return numeric measurements, the display controls it offers may not be as comprehensive as those on a DSO. In addition, the space constraints in manufacturing test systems tend to require compact instrument enclosures that preclude DSO-like timebase and amplitude controls.

Often, an engineer or technician must observe a number of measurements simultaneously. For example, pulse droop, overshoot, and average power are often monitored over the dynamic range of a test signal. For this reason, most sampling power meters offer a configurable 1-, 2-, 3-, or 4-line numerical display.

Meters that offer a live trace display (DSO-like waveform display) may also have X and Y scaling in the same manner as on a DSO (e.g., the HP8990/1—now discontinued—or the Boonton 4500A). However, this meter is more likely to have a simplified scaling method tied more closely to the requirements of peak-mode measurements. DSO display controls may be recast as shown in **Table 6** to enable adequate power-envelope trace display.

Triggering for a sampling power meter would be expected to have the following features in addition to the minimum offered by a DSO.

- Internal (level), general-purpose interface bus (GPIB), or external triggering.
- Trigger-out port.
- Three acquisition modes.
- Accuracy.
- Trigger accuracy is an important facet of the power measurement.
- Trigger accuracy should be equivalent to the measurement path.
- Linearity, temperature, and frequency are compensated to obtain the power-level comparison.
- Range.
- Settable in absolute-power value.
- Extended beyond the displayed range.
- Independent of display.

Note that free-run acquisition is distinct for the sampling power meter since there is no attempt to find and synchronize to a trigger.

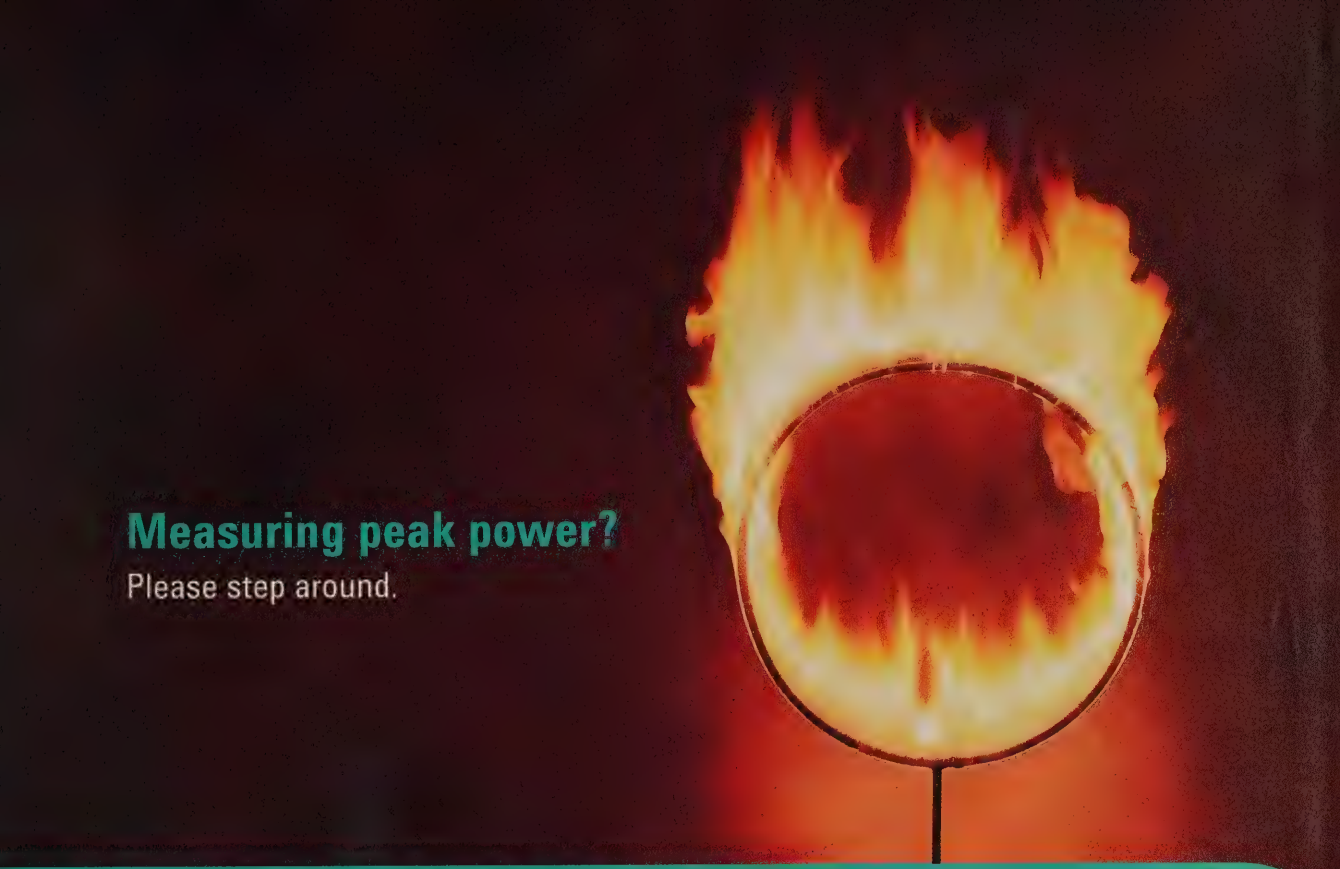
What measurements do users of a sampling power meter want to perform? Basically, there are two: peak (envelope) power or average power. When measuring power, the user may wish to define where the power is measured. The user may want more than one metric from the power envelope. So now there are two measurement categories: triggered (i.e., time relative) and nontriggered (i.e., not time relative). Since the sampling power meter is a measurement instrument, the sampled power envelope is not always required for display or inspection. The user may wish only that the sampling power meter returns a specific measurement. In power measurements, users are also interested in combinations of power. For example, the user may wish to know how much the average power changes from one pulse to the next, or from the start of the pulse to the end. The user of a sampling power meter will typically require some or all of the measurements listed in **Table 7**.

Using a fast-responding diode sensor allows the sampling power meter to track power-envelope variations and return measurements quickly. Unfortunately, diodes deliver a voltage that is proportional to power only over a portion of their operating range **(Fig. 7)**. With an input power below −20 dBm (depending on the load), a diode will deliver a voltage with a linear conversion factor from watts to volts. Above this level, the conversion factor reduces and changes from being proportional to power to being proportional to voltage. The power-to-voltage conversion law is not only dependent on the input power but is also sensitive to those changes regarding temperature and RF frequency.

To enable the sampling power meter to measure power with a diode sensor, each sample of the diode output volt-

Table 6: Display controls and counterparts on a sampling power meter

DSO		SAMPLING POWER METER		
Control	Units	Control	Units	Description
Time base	Seconds per division	Trace start	Seconds	Controls the start time of the trace relative to the trigger
Trigger position	Left, center, right	Trace length	Seconds	Controls the time span of the displayed waveform
Trigger delay	Seconds			
Amplitude range	Volts per division	Trace maximum	Watts/dBm	Sets the top of the trace
Input offset	Volts	Trace minimum	Watts/dBm	Sets the bottom of the trace



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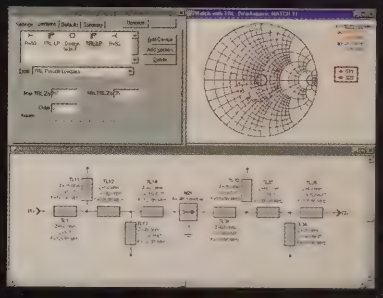

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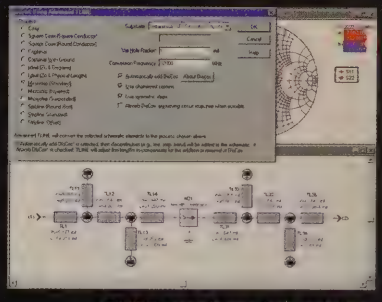

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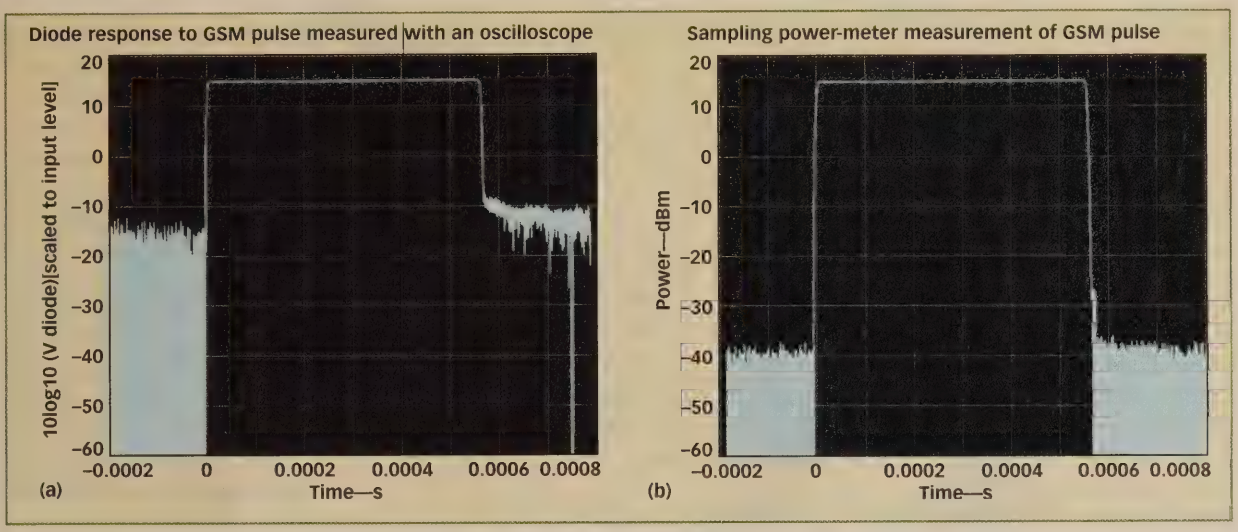

8. This is a diode signal as it appears at the ADC input, but measured with an oscilloscope (a). The same signal is measured with a sampling power meter (b).

age must be converted to power. This is achieved by sensor characterization during its manufacture. Each sensor is subjected to known power inputs and the output voltage is saved. This provides the power-to-voltage conversion characteristic of the particular sensor. Within the sampling power meter, this characteristic is converted to a voltage-to-power conversion law and applied to each input sample.

As mentioned earlier, the diode's

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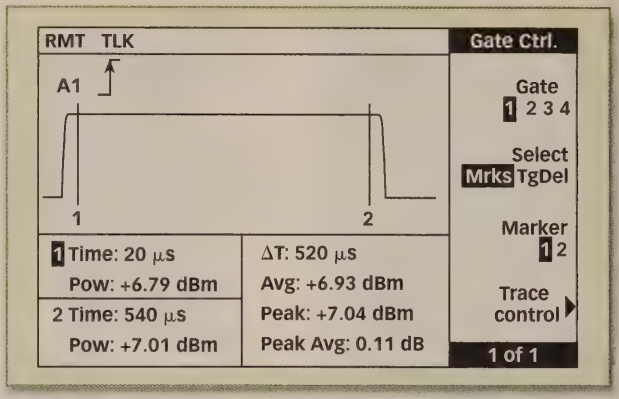
power-to-voltage conversion characteristic is highly dependent on frequency and temperature. Sensor characterization against these variables is also carried out during manufacture, and the voltage-to-power conversion law is modified when either of these variables changes. A thermistor that is mounted on the diode microcircuit, or (with less effectiveness) on the sensor body, senses temperature and the user must enter the frequency of the RF input. Compared with continuous-wave (CW) sensors, high-bandwidth peak-power sensors are particularly sensitive to changes in frequency. Thus, entering the RF frequency is very important.



9. This shows the marker setup for GSM burst-power measurement using the Agilent EPM-P.

In addition to compensating for the diode characteristics, it is also necessary to provide a frame of reference for the measurements. The corrections for the diode characteristics are intended to ensure a linear response to power, but this linear characteristic is not referred to any absolute value. To perform this, the meter normally employs a "zero & cal" procedure. "Zeroing" involves measuring any residual offset when no power is applied (the "zero" power), and subtracting this offset from any subsequent measurement. "Cal" measures the effective gain with a known power (usually 0 dBm at 50 MHz), thus providing the meter with a second reference point for the linearized characteristics and supporting absolute-power measurements.

Using sampling power meters with logarithmic scaling highlights a further distinction between these instruments and an oscilloscope. Dynamic range, when perceived in logarithmic terms, reveals the limitation of the acquisition system. A high-precision analog-to-digital converter (ADC) can compensate, but this alone cannot fully meet the requirements of some digital communications formats. For exam-

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ple, the Global System for Mobile Communications (GSM) standard calls for a ratio of burst on-to-off power of 55 dB. This would equate to 110 dB of voltage range for a truly linear sensor. Luckily, the diode's nonlinearity is advantageous here since it reduces the voltage range. However, the required range is beyond the range available from off-the-shelf ADC devices.

Some form of range switching must be provided to achieve the required dynamic range. Measuring the on-to-off ratio implies that this switching must be performed very quickly. The Agilent EPM-P power meter and E9320 peak-power sensor family offer four ranges—two ranges in the meter and two in the sensor. Sensor-range switching takes a few microseconds, while the meter ranging is indiscernible to the user. **Figure 8** shows this capability and the contrast in dynamic range compared to an oscilloscope.

Due to the power meter's nature as a sensitive measurement instrument, noise must be limited at every opportunity. This implies that the bandwidth be restricted to the minimum required to support the measurement. Nyquist sampling can be used to limit the bandwidth to the minimum required, but an anti-alias filter must be used before the sampler to prevent aliasing. Neither of these are strictly necessary in all circumstances, but not following Nyquist forces some other assumptions on the signal. To achieve good time-domain response, one must restrict the passband flatness available from an analog filter. Good time-domain response is needed to avoid overshoot on fast-rising edges. However, continuous signals with bandwidth less than the anti-alias filter will suffer degraded peak measurement due to the filter's passband droop.

Digital-signal-processing (DSP) techniques can be used to compensate for the droop. They can also be used to reduce the measurement bandwidth to cover the input signal and, thus, provide some additional noise-reduction benefit. This can be switched off when fast pulses are being measured. Compensation filters are generated by measuring the frequency response of the anti-alias filter, calculating the complementary response to this, and combining that with a filter designed to restrict the bandwidth.

Continuous sampling has advantages in the amount of noise bandwidth for a particular measurement bandwidth. It has another advantage that is useful for examining modern digital-modulation formats. Since the measured signal can be reconstructed perfectly, it is possible to examine modulation artifacts from random data signals. An example of this may be residual amplitude modulation (AM) in constant-AM schemes such as Gaussian minimum-shift keying (GMSK). If examined with a random sampler, it would be impossible to tell random data from noise, whereas with continuous sampling it is easy to distinguish between the two. Random sampling relies on repetition of the signal to build up a picture of the signal, but data signals do not repeat. So each new capture has no relation to the last and, hence, appears to be noise. On-off events and the signal before and

Continued on page 122

Table 7: Sampling power-meter measurements

MEASUREMENT		UNITS	DESCRIPTION
Triggered	Average	Watts/dBm	The average power of the waveform displayed on the screen
	Peak		The peak power of the waveform displayed on the screen
	Pulse top		The power at the top of the pulse signal (the most likely voltage at the pulse top)
	Overshoot	Percent	
	Peak-to-average ratio	Percent/dB	
	Rise time	Seconds	The time it takes for a signal to rise from 10 to 90 percent of the pulse height
	Fall time		The time it takes for a signal to fall from 90 to 10 percent of the pulse height
	Period		The time required for one full repetition of the waveform
	Frequency	Hertz	The frequency of the signal (the reciprocal of period)
	Duty cycle	Percent	The ratio of high-to-low-pulse intervals for a pulsed waveform
Non-triggered	Average	Watts/dBm	The average of the RF power at the sensor input
	Peak		The peak of the RF power at the sensor input
	Peak-to-average ratio	Percent/dB	
Gated	Average	Watts/dBm	The average power of the waveform during a time interval defined by the measurement gate
	Peak		The peak power of the waveform during a time interval defined by the measurement gate
	Peak-to-average ratio	Percent/dB	The peak-to-average ratio of the waveform during time interval defined by the measurement gate
Combination	Gates	Watts/dBm/ percent/dB	Add/subtract/multiply/divide combination of gated measurements from two different gates
	Channels		Add/subtract/multiply/divide combination of measurements from two different channels

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By following these design rules and straightforward techniques, it should be possible to create small-signal LNAs that can be used under a variety of conditions.

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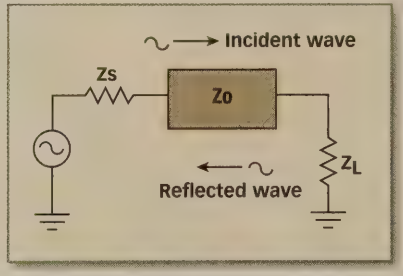
Designing low-noise amplifiers (LNAs) usually follows a procedure according to a specific set of operating conditions. Unfortunately, some designers may not choose the bias supply or other conditions outlined in the design guidelines. This three-part article series is meant to help engineers who must develop circuits for a particular set of operating conditions. This first part reviews transmission lines and RF power-gain definitions.

Voltage, current, or power emanating from a source impedance, Z_S, and delivered to a load impedance, Z_L, can be regarded as the sum of incident and reflected waves traveling in opposite directions along a transmission line **(Fig. 1)** of characteristic impedance Z_0. If Z_L equals Z_0 exactly, the incident wave is totally absorbed in the load and produces no reflected wave.

If Z_L differs from Z_0, some of the incident wave is not absorbed in the load, but is reflected back toward the source. If the source impedance Z_S equals Z_0, the reflected wave from the load is absorbed in the source and no further reflection occurs. For the case where Z_S is not equal to Z_0, a portion of the reflected wave from the load is re-reflected from the source back toward the load. For a lossless transmission line, this process repeats indefinitely.



1. The amount of mismatch between a transmission line with characteristic impedance of Z_0 and a source or load impedance determines the amount of incident energy that is reflected.

The degree of mismatch between Z_0 and Z_L (or Z_S) determines the amount of incident wave that is reflected. The ratio of reflected wave to incident wave is known as the reflection coefficient, and is simply a measure of the quality of the match between the transmission line and the terminating impedance. The reflection coefficient, Γ, is a complex quantity expressed in polar form as a magnitude and an angle (Fig. 1):

$$\Gamma = WAVE_{reflected} / WAVE_{incident} = \rho / \theta \qquad (1)$$

As matching between the characteristic impedance of the transmission line and the terminating impedance improves, the reflected wave becomes smaller and the reflection coefficient

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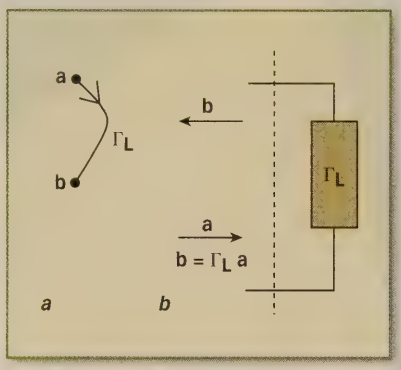


2. A flow graph (a) helps in the visualization of the load reflection coefficient as a transmission line (b).

decreases. With a perfect match, there is no reflected wave and the reflection coefficient equals zero. If the load impedance is an open or short circuit, no incident power can be absorbed in the load, and all of the power must be reflected back toward the source. In that case, the reflection coefficient is equal to one. The normal range of magnitude for the reflection coefficient is between zero and one.

For reflection coefficients that are greater than one, the magnitude of the wave that is reflected from the load impedance should be greater than that of the incident wave to that load. It follows, therefore, that the load in question must be a source of power. This concept is useful in designing an oscillator, but at the input network of an amplifier, it represents instability. Reflection coefficients can be expressed in terms of the impedances under consideration. For example, the reflection coefficient at the load can be expressed as:

$$\Gamma = (Z_L - Z_0)/(Z_L + Z_0) \text{ or with}$$
$$Z_L = Z_L / Z_0 \text{ in a normalized form}$$
$$\Gamma = (Z_L - 1)/(Z_L + 1) \qquad (2)$$

For impedance matching in microwave and RF networks, the source and load impedances are often expressed in terms of the source reflection coefficient, Γ_S, and load reflection coefficient, Γ_L. Incident and reflected waves are represented by wave flow graphs. Flow graphs allow engineers to visualize linear relations that are among different variables in a network, and they help in rapidly constructing a transfer function that is between two points in the network **(Fig. 2a)**.

Each variable is represented by a node in the graph, and the different nodes are linked together by directive paths that provide the relations between related and unrelated variables. Paths are directed from the node representing the independent variable to those representing the dependent variables, and each path is assigned a gain that is related to the reflection coefficient that links the two variables **(Fig. 2b)**.

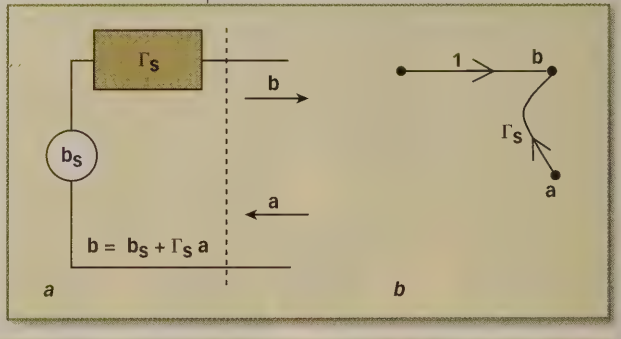
In **Fig. 3**, variables a and b are complex values that are associated with the incident and reflected waves, respectively. The presence of variable b depends on the presence of variable a. The source reflection coefficient can be expressed as b/a related to the source power when b_S is zero. When the power source is matched 100 percent to the transmission line, $b_S = b$.

By inserting a two-port network between source and load in the circuit in the left side of Fig. 3, the diagram of **Fig. 4** results. In Fig. 4, S_{11} is the input reflection coefficient, S_{12} is the reverse transmission coefficient, S_{21} is the forward transmission coefficient, S_{22} is the output reflection coefficient, and $\Delta S = S_{11}S_{22} - S_{21}S_{12}$.

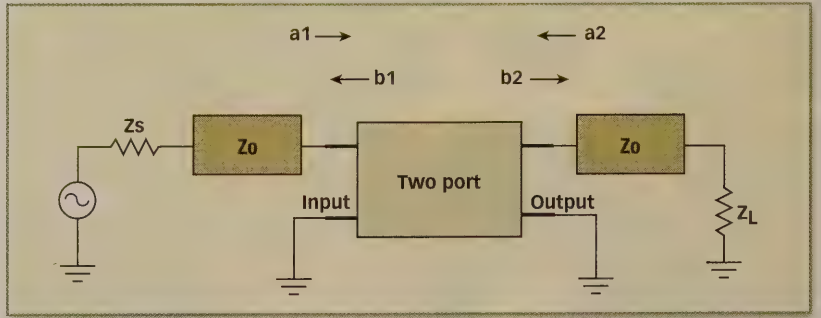
With reference to Fig. 4, for any traveling wave that originates at the source:

- A portion of the wave that originates at the source and is incident on the two-port device (a_1) will be reflected (b_1), while another portion will be transmitted through the two-port device.
- A fraction of the transmitted signal is then reflected from the load and becomes incident on the output port of the two-port device (a_2).
- A portion of the signal (a_2) is then reflected from the output port back toward the load (b_2), while a fraction is transmitted through the two-port device to the source.

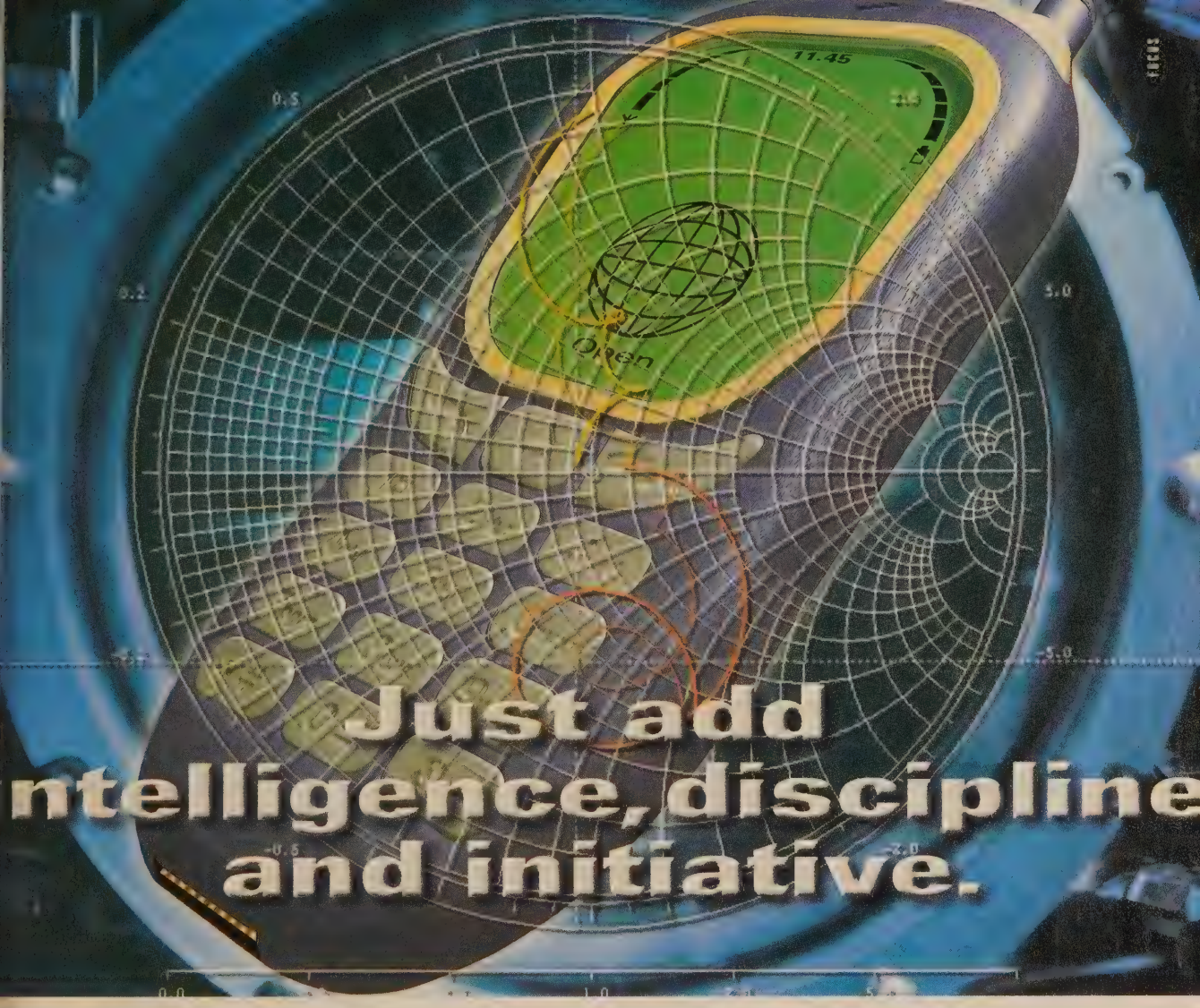
Any traveling wave that is present in the circuit is composed of two components. For instance, the traveling-wave components that flow from the output of the two-port device to the load consist of the portion of a_2 that is reflected from the output of the two-port device, and the portion of a_1 that is transmitted through the two-port device. Similarly, the traveling wave that flows from the input of the two-port device back toward the source consists of the



3. The power-source reflection coefficient can also be visualized with a block diagram (a) and a flow graph (b).



4. A simple two-port network diagram is useful for graphing S-parameters.



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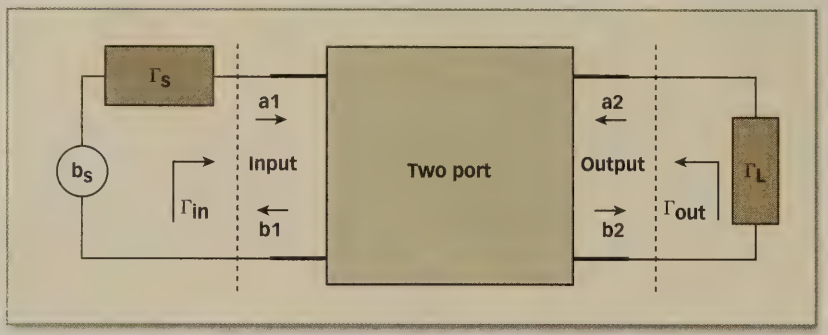


5. The two-port network diagram has been enhanced for the available gain definition.

portion of a_1 that is reflected from the input and the fraction of a_2 that is transmitted through the two-port device. These observations in equation form to appear as:

$b_1 = S_{11}a_1 + S_{12}a_2$ and $b_2 = S_{21}a_1 + S_{22}a_2$. If a_2 is set to zero, then:

$S_{11} = 1/2b_1/a_1 1/2a_2 = 0$, and if a_1 is set to zero, then:

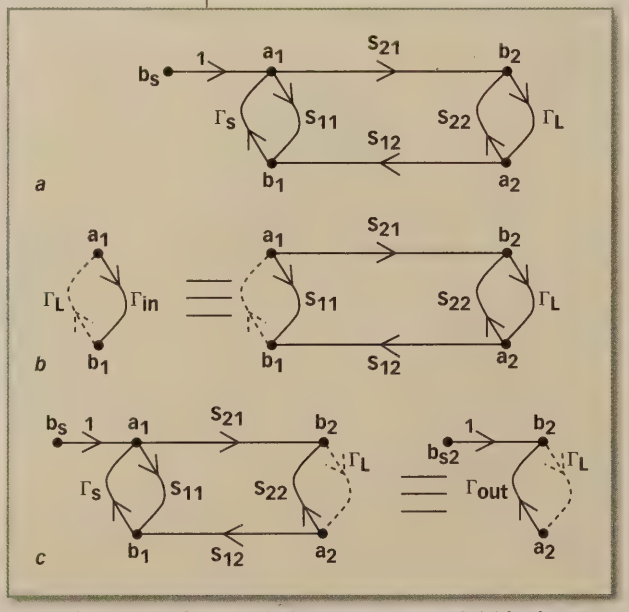
$S_{22} = 1/2b_2/a_2 1/2a_1 = 0$.

By definition, a reflected wave that is divided by an incident wave is equal to the reflection coefficient. A two-port network (such as the LNA) can be characterized completely by its scattering parameters (S-parameters).

With S-parameters, it is possible to calculate potential instabilities (an amplifier's tendency to oscillate), maximum available gain, input and output impedances, and transducer gain. It is also possible to calculate the optimum source and load impedances, either for simultaneous conjugate matching, or simply to help choose specific source and load impedances for a specified transducer gain.

Figure 5 shows a two-port network connection for the available gain definition, where Γ_{in} is the two-port input reflection coefficient and the output is terminated in Γ_L. Parameter Γ_{out} is the two-port output reflection coefficient, as long as the input is terminated with Γ_S.

The flow graph for a loaded two-port network **(Fig. 6a)** shows a closed loop that allows starting from one node and returning to that node without going through it twice. The precise physical significance of these loops is illustrated by a two-port network, such as an LNA, whose output is closed on a load



6. The flow graph for a loaded two-port network (a) is shown with output closed on a load Γ_L (b) and closed on a source impedance (c).

Γ_L **(Fig. 6b)**.

In order to define the wave b_1 reflected back to the input, one must follow the different paths that lead from a_1 to b_1. The possible paths are S_{11}, $S_{21}\Gamma_L S_{22}\Gamma_L S_{12}$, $S_{21}\Gamma_L S_{22}\Gamma_L S_{22}\Gamma_L S_{12}$, and so on.

It appears that the loop $S_{22}\Gamma_2$ represents a multiple reflection due to the two-port network reflection coefficient between the output and the mismatched load, with:

$$b_1 = S_{11}a_1 + S_{21}\Gamma_L S_{12}a_1 1 + S_{22}\Gamma_L + \ldots + (S_{22}\Gamma_L{}^n + \ldots.)$$

where:

the bracketed terms = the multiple reflection:

$$1/(1 - S_{22}\Gamma_L)$$

The network input reflection coefficient for the two ports is:

$$\Gamma_{in} = b_1 / a_1 = S_{11} + (S_{21}\,\Gamma_L\,S_{12}) / (1 - S_{22}\,\Gamma_L) \quad (3)$$

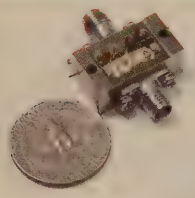
When the two-port network input is closed on a source impedance, the reflection coefficient that is seen from the output is **(Fig. 6c)**:

$$\Gamma_{out} = b_2 / a_2 = S_{22} + (S_{21}\,\Gamma_S\,S_{12}) / (1 - S_{11}\,\Gamma_S) \quad (4)$$

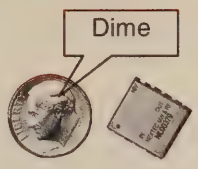
Waves a and b are directly related to the concept of power.

Part 2 of this series will continue with a review of transmission lines and RF theory by using power-gain definitions to develop the mathematical concept of maximum available gain (MAG) in a two-port amplifier network. Part 3 explores amplifier stability (the tendency for oscillation), impedance matching, and general amplifier design, using the S-parameters as design tools. Part 4 will close out the series with some practical design examples, including how to match an LNA for MAG, as well as how to match an LNA for particular gain performance. MRF

NEW ! Small Enough, Good Enough



Broadband Amplifiers (2-8 GHz | 6-18 GHz | 18-26.5 GHz | 26-40 GHz) have been modified to have a smaller size. Four models (NB00375, NB00376, NB00377 and NB00378) are available now. Each model covers 2-8 GHz, 6-18 GHz, 18-26.5 GHz and 26-40 GHz, respectively. Typical gains of those models are around 20 dB and output powers are about 20 dBm.



New Tiny LNA's (NF: 1.1 dB, 1810-1910 MHz | 1710-1780 MHz) have been developed. Four models (NL00379, NL00380, NL00381 and NL00382) are available now with Noise Figure as low as 1.1 dB. Two of them covers 1850-1910 MHz and the other two models cover 1710-1780 MHz. Typical gain is 25 dB. Output IP3 are 33 - 39 dBm. **Two more models** will come out in October 2001 with higher gain of 36 dB covering 1850-1910 MHz.

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Read a menu for flexible-circuit choices

FLEXIBLE CIRCUITS HAVE been a part of the electronic equipment landscape for more than 25 years. But the latest types, the materials they are constructed of, and the applications they serve are not an everyday consideration for most hardware designers. To get a broad overview of what is available in the world of flexible circuits, go to a website run by Rogers Corp.'s Advanced Circuit Materials Division (Chandler, AZ). Here, the company's Flex Circuit families are broken down in a menu-like way, allowing one to make a top-level selection and proceed down to the detailed data-sheet level if necessary.

For example, if "Recommendation by Application" is selected, a user quickly sees that applications are divided into 10 categories, such as general purpose, hard-disk drive, cell phone, automotive, and more. Selecting the cell-phone category allows a web surfer to examine topics such as, "Where Flex Circuits are Used," "Material Considerations," and "Product Recommendations." Under the where-they-are-used section, the four main cell-phone applications are stated—interconnect between the main board and other modules, dynamic use in which the Flex Circuit is used in a hinge area where the microphone folds out in order to connect the cell-phone battery to the main board, etc. From here, a user can probe deeper to choose either Material Considerations or Product Recommendations.

Material Considerations provides a checklist of the basic questions that must be asked prior to selecting a flexible circuit for a specific application (cost, adequate flex, UL rating, current carrying capacity, etc.). Product Recommendations provides information about the various types of Flex Circuits that are manufactured by Rogers, which is a basic description of each different Flex Circuit product line. From here, the complete data sheets can be accessed, along with a Processing Guidelines page that discusses considerations for plating, drilling, and other manufacturing factors involved with flexible circuits.

Rogers Corp. Advanced Circuit Materials Div., Chandler, AZ; Internet: www.rogers-corp.com/cmu/divproducts.htm

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The CD-ROM was an excellent vehicle for demonstrating how time-domain-reflectometry measurements are made.

Looking at time-domain-reflectometry theory

TIME-DOMAIN REFLECTOMETRY is a useful design tool for understanding how circuits and interconnects on a PCB behave at gigabit-per-second data rates. With digital data rates already at gigabit speed, it is increasingly important that designers know how to optimize components, such as microstrip lines, PCB traces, coaxial cables, SMA connectors, etc. to reduce or eliminate distortions and reflections. The success or failure of a high-speed design depends on optimizing components.

Two years ago, the Hewlett-Packard Test and Measurement Division—spun off and now part of Agilent Technologies (Palo Alto, CA)—released a CD-ROM entitled "Time Domain Reflectometry with the HP54750A Digitizing Oscilloscope." The CD-ROM was an excellent vehicle for explaining the theory of time- domain reflectometry and demonstrating how time-domain-reflectometry measurements are made using an animated format. The basic idea of time-domain-reflectometry measurements is to launch a fast rise-time edge into a DUT, such as a transmission line, and monitor the reflections with a high-bandwidth oscilloscope. On the CD-ROM, a user views animated incident and reflected waves of a step generator that launches a signal into a transmission line and how the reflected signal would appear on the oscilloscope. The animation is performed for four different conditions: series resistance, shunt and series capacitance, and series inductance. Unfortunately, the time-domain-reflectometry CD-ROM is not available on Agilent's current website, but a host of application notes and technical papers on the subject is available for downloading from the company's website. Two application notes that might be of interest to many designers are "TDR Theory" (No. 1304-2) and "Evaluating Microstrip With TDR" (No. 1304-1). Agilent also has a time-domain-reflectometry probe, the N1020A, which works with their 83480A communications analyzer—a replacement for the HP54750A that was the demonstration instrument in the CD-ROM. The 54754A differential time-domain-reflectometry plug-in module from the original demonstration is still available from the company.

Agilent Technologies, Test and Measurement Organization, Palo Alto, CA 94303-9511; Internet: www.agilent.com/cm/rdmfg.

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cover story

SiGe Direct Modulators Drive BTS Designs To 4 GHz

This pair of direct quadrature modulators allows designers to eliminate a frequency-conversion stage in their BTS designs without sacrificing dynamic range.

THOMAS CAMERON
Manager, RF Module Design
SARA MESZAROS
Design Engineer
Stanford Microdevices, Inc., 522 Almanor Ave., Sunnyvale, CA 94086; (800) 764-6642, (408) 616-5400, FAX: (408) 739-0970, Internet: www.stanfordmicro.com.

Silicon germanium (SiGe) has held the promise of wide dynamic range and linearity for some time. A pair of direct quadrature modulators from Stanford Microdevices (Sunnyvale, CA) makes good on that promise by supporting the design of simpler wireless base-transceiver stations (BTS), while delivering industry-leading dynamic range and linearity. The modulators are the model STQ-2016 for mobile wireless applications from 700 to 2500 MHz and the model STQ-3016 for use in fixed wireless systems from 2500 to 4500 MHz.

A direct modulator can be used to produce virtually any analog or digital modulation scheme, including amplitude, phase, and frequency modulation (FM). With adequate bandwidth, the modulator can support a wide range of modulation schemes. In contrast to the conventional dual-conversion modulator approach, direct modulators eliminate a frequency-conversion stage [filters, RF mixer, and a local oscillator (LO)], while also reducing the size, cost, and complexity of base-station transceivers.

In comparison to a typical dual-conversion approach, a direct-conversion architecture eliminates the need for the filters, RF mixer, and one LO. However, to implement a direct-modulation architecture, the modulator must be capable of very-high performance. For example, it should have sufficient carrier and sideband suppression, and a broadband noise floor that exceeds −150 dBm/Hz.

Both of the new modulators are designed for use with a single +5-VDC supply. The lower-frequency STQ-2016 is suitable for Global System for Mobile Communications (GSM), cellular, and personal-communications-

services (PCS) bands from 700 to 2500 MHz, while the STQ-3016 is the first direct modulator for fixed wireless applications from 2500 to 4000 MHz.

Both integrated circuits (ICs) contain the entire quadrature (Q)-modulation function on a single device. Balanced baseband inputs are modulated in Q and combined in a summing amplifier. Fast shutdown capability in the summing amplifier provides 60-dB shutdown attenuation. Each direct-modulator chip includes LO buffer amplifiers which provide gain in the LO path and isolation between the in-phase (I) and Q paths. The LO drive requirement is low, at −5 dBm. Q phase shift is accomplished using an on-chip polyphase network. The modulators' baseband ports feature a 3-dB bandwidth exceeding 500 MHz that is measured in a 50-Ω system, making them suitable for a wide range of wireless systems.

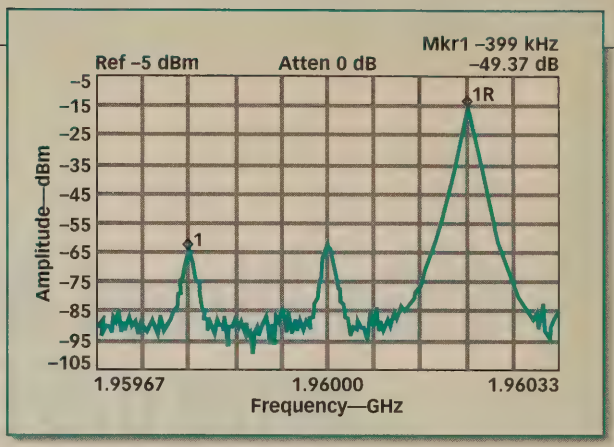
The entire modulator design is based on a balanced configuration from baseband to RF to optimize linearity and reject spurious signals. The monolithic approach delivers carrier suppression of −40 dBm, typically with a LO input of −5 dBm. Printed-circuit board (PCB) and hybrid-modulator designs that are assembled from discrete mixers, couplers, and gain blocks cannot achieve more than −20 to −25-dBm carrier suppression without expensive tuning.

Sideband suppression is −40 dBc for the STQ-2016 and −33 dBc for the STQ-3016. Carrier suppression may be maximized further by using an external DC offset circuit that is easily implemented with operational amplifiers. Another important feature of the STQ family is the exceptionally low broadband noise floor. Typically, the noise floor is better than −154 dBm/Hz for the STQ-2016 and 153 dBm/Hz for the STQ-3016.

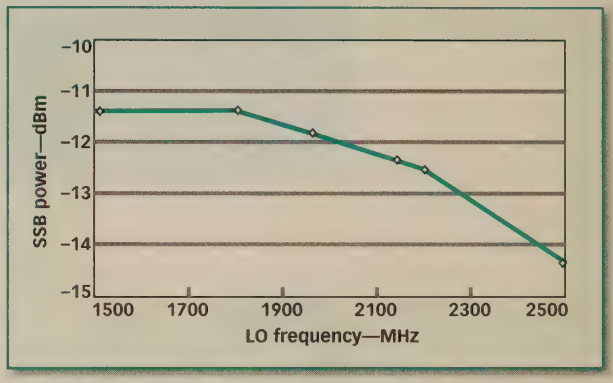
The STQ-2016 has output power at 1-dB compression of +3 dBm, allowing the device to be linearly operated at an output power of approximately −10 dBm, depending on the modulation scheme that is being employed. The STQ-3016 has an output power at 1-dB compression of +1 dBm, which should be adequate for the higher-linearity modulation schemes that are being implemented in fixed wireless, such as high-level Q-amplitude modulation (QAM) and offset-frequency-division-multiplex (OFDM) modulation.

A combination of signal sources was used to test the modulators. A model 8904A multifunction synthesizer from Agilent Technologies (Santa Rosa, CA) provided low-frequency single-tone baseband signals, while a model E4437B digital signal generator from the same supplier was used for higher-frequency modulated signals, such as wideband code-division multiple access (WCDMA). These generators are capable of producing quadrature baseband signals. For testing, baseband signals are split into balanced signals and boosted by operational amplifiers.

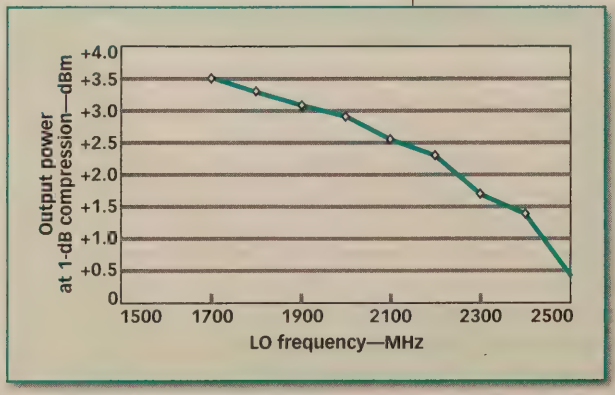
Carrier signals can be created with any high-performance signal generator. The test source should have low phase



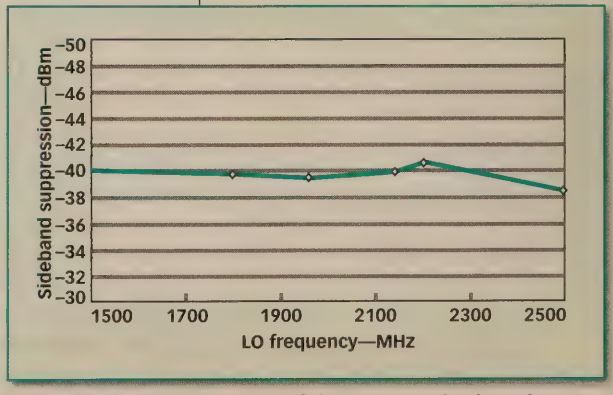
1. The SSB output spectrum of the STQ-2016 was measured at 1.96 GHz using 200-kHz quadrature tones.



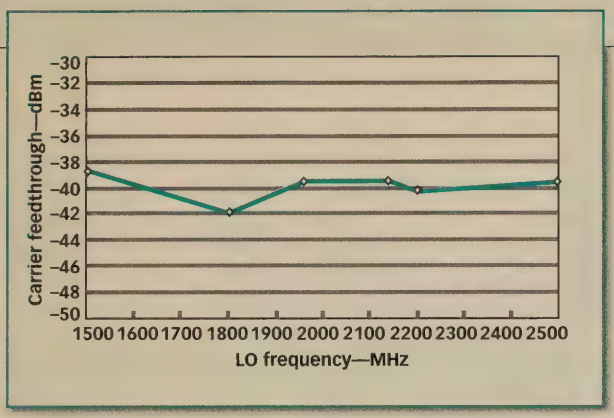
2. The SSB output power of the STQ-2016 is plotted versus LO frequency.



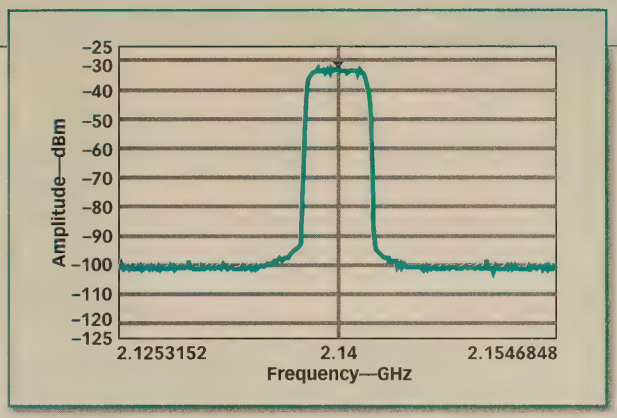
3. The output power at 1-dB compression for the STQ-2016 is plotted as a function of LO frequency.



4. The sideband suppression of the STQ-2016 is plotted as a function of LO frequency.



5. The carrier feedthrough is plotted as a function of LO frequency for the STQ-2016 modulator.



6. The ACPR for the STQ-2016 in the UMTS band was measured using WCDMA 3GPP modulation.

noise and low harmonics so that the LO does not affect the measurements. For testing the two modulators, the model 4437B signal generator was used to generate the carrier signals. Single-ended carrier signals were fed to a test fixture and converted on board the fixture to balanced signals to drive the balanced LO input ports. Balanced RF outputs feed a balun which converts them back to single-ended signals. The RF output is matched to the band of interest and measured with a spectrum analyzer (with front-end preamplifier to improve noise measurements).

The generally accepted method of evaluating a modulator is known as single-sideband-suppressed-carrier (SSBSC) modulation. SSB modulation enables a standard evaluation of some of the key characteristics of a modulator, such as linearity, carrier suppression, and sideband suppression. Q-sinusoidal-baseband tones are injected at a frequency of interest and a sinusoid is applied to the LO port at the desired carrier frequency. Measurements will reveal the spectrum at the carrier frequency **(Fig. 1)**. In this case, 200-kHz tones were applied to the baseband ports at an amplitude of 600-mV peak-to-peak and with a −3-dBm carrier at 1960 MHz. The three tones from right to left are the desired modulated tone, the suppressed carrier, and the suppressed sideband signal. In this case, the sup-

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pressed carrier is at −58 dBm, which can be improved by applying a DC offset. The sideband suppression, which is −49 dBc, is an indication of the good balance achieved between the I and Q paths on the chip.

Figure 2 shows SSB power over the range of LO frequencies. These data are taken by sweeping the LO with a fixed baseband frequency. The output power is flat up to 2 GHz, dropping at the higher end of the band. It is interesting to note that the output power is extremely flat over any 60- or 75-MHz band in this range, such as the PCS or digital-communications-services (DCS) bands.

Figure 3 shows output power at 1-dB compression versus LO frequency. The output power at 1-dB compression is +3.5 dBm at 1700 MHz, dropping off to +0.5 dBm at 2500 MHz. **Figure 4** shows sideband suppression versus LO frequency and **Fig. 5** plots carrier feedthrough versus LO frequency, showing excellent sideband suppression and carrier feedthrough suppression over the entire range.

While the previous data provide a good figure of merit for a modulator, they are applicable to narrowband modulation. For wideband-modulation formats, such as CDMA and WCDMA, the adjacent-channel power ratio (ACPR) is a good figure of merit for the dynamic range of the modulator.

The test setup for ACPR measurements is similar to the previous tests, the only difference being the baseband function generator, an Agilent E4437B digital signal generator. The E4437B has built-in functions that are based on industry standards. For these tests, the built-in WCDMA Third Generation Partnership Project (3GPP) test model (No. 2) with three dedicated physical channels (DPCHs) was used to generate the baseband signal. The LO frequency was set at 2.14 GHz, which is the Universal Mobile Telecommunications Services (UMTS) standard frequency.

The ACPR measurements reveal impressive performance levels for the STQ-2016 **(Fig. 6)**. With channel power of −13.19 dBm, the measured upper ACPR is −66.14 dB while the lower ACPR is −65.79 dB.

The STQ-2016 and STQ-3016 direct-Q modulators should greatly simplify the design of wireless base-station transceivers. The STQ-2016 is suitable for most mobile wireless applications, while the STQ-3016 will support fixed wireless applications from 2.5 to 4.0 GHz. **Stanford Microdevices, Inc., 522 Almanor Ave., Sunnyvale, CA 94086; (800) 764-6642, (408) 616-5400, FAX: (408) 739-0970, Internet: www.stanfordmicro.com.**

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PRODUCT technology

Multilayer Magic Yields Shrinking Circuits

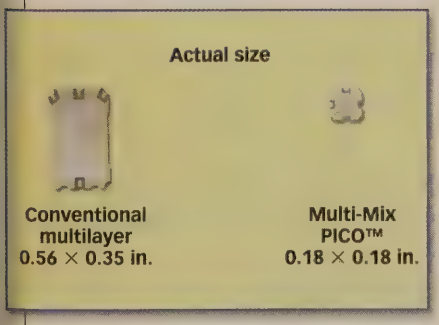
Multi-Mix PICO reduces the size of components, such as quadrature hybrids and directional couplers, by 84 percent when compared to any conventional competitive product.

JACK BROWNE
Publisher/Editor

Multi-Mix multilayer technology burst on the market in 1998, with components comparable in size to the smallest of those then available, and performance competitive with any other technology. Now, three years later, Merrimac Industries (West Caldwell, NJ) has introduced the next generation of Multi-Mix, known as Multi-Mix PICO, along with a family of quadrature hybrids and directional couplers (the Z series) that is up to 87 percent smaller than its predecessors, while still handling 100-W continuous-wave (CW) power with appropriate heat sinking.

The small footprint realized by the Multi-Mix PICO components is dramatic when they are compared to the footprint of their predecessors **(Fig. 1)**. The size reduction should be well-received by designers of two-and-a-half-generation (2.5G) and third-generation (3G) wireless base stations, picocells, multicarrier amplifiers, and other wireless systems, where space is increasingly at a premium. All of the components have a standard footprint and are based on a library of Multi-Mix configurations, which simplifies the design process and allows the company to offer them as off-the-shelf components with delivery from stock.

Multi-Mix is a unique circuit-design and fabrication process that allows microwave circuits to be constructed in three dimensions (3D) rather than in the conventional two-dimensional (2D) planar format. It provides tremendous advantages in size, weight, shield-



1. The Multi-Mix PICO footprint is compared to its predecessor, which was formerly the smallest footprint available.

Table 1: Multi-Mix PICO quad hybrids

Parameter	QHD-22-1.9G	QHD-22-2.1G
Frequency range	1.7 to 2.0 GHz	2.0 to 2.3 GHz
Insertion loss	0.35 dB (max)	0.35 dB (max)
Isolation		
Paths 1 to 3	20 dB (min)	20 dB (min)
Paths 2 to 4	15 dB (min)	15 dB (min)
VSWR		
Ports 2 and 4	1.30:1 (max)	1.40:1 (max)
Ports 1 and 3	1.50:1 max)	1.60:1 (max)
Amplitude imbalance	±0.4 dB (max)	±0.3 dB (max)
Phase imbalance	±3 deg. (max)	±4 deg. (max)
Maximum input power	100 W CW (max)	100 W CW (max)
Size	0.18 × 0.18 × 0.09 in.	
Weight	0.003 oz.	0.003 oz.

ing effectiveness, and electrical performance compared to conventional stripline approaches. The Multi-Mix circuit consists of various circuit and isolation layers. The layers are fusion-bonded together, with multiple layers aligned through indexing holes. Plated-through holes are used to provide electrical and ground connections between layers. The fusion-bonding process delivers a multilayered, homogenous monolithic structure that is extremely compact, and supports integration of active and passive components. The company has manufactured a considerable number of Multi-Mix multifunction modules, including filter banks. Multi-Mix PICO will allow these same modules to be manufactured with significantly smaller dimensions.

The Multi-Mix PICO components include the Z series of quad hybrids and directional couplers. The quad hybrids include the 1.7-to-2.0-GHz model QHD-2Z-1.9G and the 2.0-to-2.3-GHz model QHD-2Z-2.1G. The tiny stripline components feature outstanding phase and amplitude balance with 0.35 dB or less insertion loss **(Table 1)**, and measure only 0.18 × 0.18 × 0.09 in. (0.46 × 0.46 × 0.23 cm) while weighing only 0.003 oz. The hybrids are fabricated on stable ceramic-filled polytetrafluoroethylene (PTFE) dielectrics that are compatible with a wide range of substrates, including FR-4, polyamide, and G-10 materials. A wraparound ground plane provides a high level of electromagnetic-interference (EMI) shielding.

The Z series of Multi-Mix PICO directional couplers includes the 10-dB models CSD-10Z-1.9G and CSD-10Z-2.1G, as well as the 20-dB CSD-20Z-2.0, all with an operating frequency range of 1.7 to 2.3 GHz. As with their quad hybrid counterparts, the directional couplers measure only 0.18 × 0.18 × 0.09 in. (0.46 × 0.46 × 0.23

Table 2: Multi-Mix PICO couplers

Parameter	CSD-10Z-1.9G	CSD-20Z-1.9G	CSD-10Z-2.1G	CSD-20G-2.0G
Frequency range	1.7 to 2.0 GHz	1.7 to 2.0 GHz	2.0 to 2.3 GHz	2.0 to 2.3 GHz
Coupling	10.5 dB	19.5 dB	10.5 dB	20.0 dB
Coupling flatness	±1.0 dB	±1.0 dB	±1.0 dB	±1.0 dB
Insertion loss	0.2 dB (max)	0.2 dB (max)	0.2 dB (max)	0.2 dB (max)
VSWR	1.30:1 (max)	1.30:1 (max)	1.30:1 (max)	1.30:1 (max)
Frequency sensitivity	±0.5 dB	±0.3 dB	±0.3 dB	±0.2 dB
Directivity	15 dB (min)	13 dB (min)	15 dB (min)	15 dB (min)
Input power (max)	100 W CW	100 W CW	100 W CW	100 W CW

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800-1000	0.5	20	0.2	2.0	1.40:1	1.0	PB-2	$4.99
1750-2050	0.5	20	0.2	2.0	1.40:1	1.0	PB-4	$4.99
2200-2500	0.6	18	0.3	3.0	1.40:1	1.0	PB-5	$4.99
800-1000	0.6	18	0.2	2.0	1.40:1	10.0	PB-14	$9.99
1300-1600	0.7	18	0.2	2.0	1.40:1	10.0	PB-10	$9.99
1700-2000	0.7	18	0.2	2.0	1.40:1	10.0	PB-11	$9.99
2300-2700	0.7	18	0.3	3.0	1.40:1	10.0	PB-12	$9.99
3000-3800	0.8	18	0.3	3.0	1.40:1	10.0	PB-13	$9.99

90° Hybrids

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Freq. Range (GHz)	Coupling Loss (dB)	Iso. (dB) min.	Amp. Balance (dB) max.	Phase Balance (Deg) max.	VSWR max.	P/N	Cost (Qty. 5-9)
5.09-5.25	3.2 + 0.2	20	0.2	2.0	1.30:1	QS2-B6-463/2	$99.99
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4.00-8.00	3.3 + 0.3	18	1.4	4.0	1.25:1	QS2-05-463/2	$99.99
6.10-6.40	3.2 + 0.2	20	0.2	2.0	1.30:1	QS2-B7-463/2	$99.99
4.50-9.00	3.2 + 0.3	18	1.0	2.0	1.30:1	QS2-B10-463/2	$99.99
10.80-12.00	3.3 + 0.3	20	0.5	2.0	1.25:1	QS2-B9-463/2	$99.99
12.50-13.50	3.3 + 0.3	20	0.5	2.0	1.25:1	QS2-B11-463/2	$99.99

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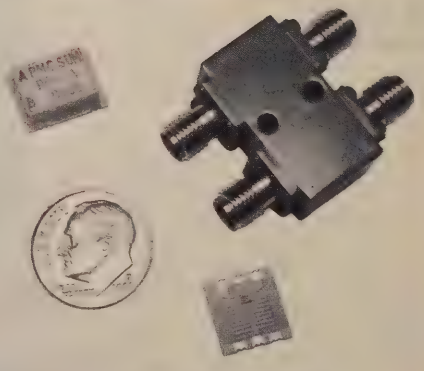

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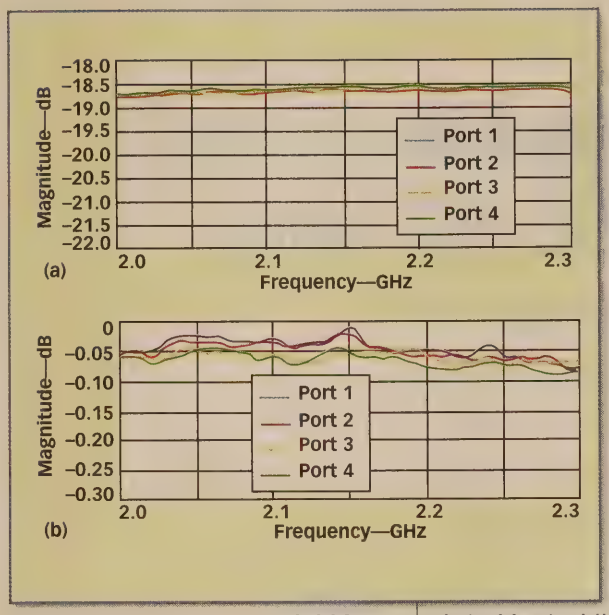

2. Measurements performed on a model CSD-20Z-2.0G directional coupler show the coupling to be flat and uniform at all ports (top), with insertion loss that is considerably less than the rated 0.2-dB specification (bottom).

cm) while weighing only 0.003 oz. The couplers have insertion loss of 2 dB, with accuracy within ±1 dB. Directivity is 13 dB or better for all models **(Table 2)** while the worst-case frequency sensitivity is ±0.3 dB. Measured data **(Fig. 2)** show the rated specifications to be somewhat conservative in terms of coupling flatness and insertion loss.

The Multi-Mix PICO directional couplers and quad hybrids are rated for operating temperatures from −55 to +85°C. The components are supplied on tape and reel for automatic assembly or can be integrated with other components as part of a Multi-Mix multifunction module. The Multi-Mix product families have passed many environmental screening tests, including thermal and mechanical shock, burn-in, acceleration, vibration, moisture resistance, solder heat resistance, and thermal cycling of more than 1000 cycles. P&A: $0.95 for standard models (250,000 qty. or less); stock. **Merrimac Industries, Inc., 41 Fairfield Pl., West Caldwell, NJ 07006; (888) 434-MMFM, FAX: (973) 882-5990, e-mail: MMFM@merrimacind.com, Internet: www.multi-mix.com.** MRF

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Low-Cost Design Kit Boasts 90 HBT Amps

This low-cost design kit features 90 packaged monolithic HBT amplifiers, extensiveS-parameter data, a connectorized test fixture, and a handy biasing chart.

JACK BROWNE
Publisher/Editor

Microwave amplifiers used to sell for about $100 each. These days, that same price will buy a model K1-GAL Designer's Kit from Mini-Circuits (Brooklyn, NY), with 90 GAL-series broadband monolithic amplifiers, design data, and a fully assembled SMA test fixture. The 90 amplifiers include 10 units each of nine different models, covering a total frequency range of DC to 8 GHz.

Based on InGaP HBT technology, the GAL-series amplifiers are housed in miniature SOT-89 plastic surface-mount packages. In addition to the amplifiers and the test fixture, the kit includes an 11-page brochure containing specifications and data for the amplifiers, a copy of the company's RF/IF Designer's Guide, and an instruction sheet on the use of the test fixture (complete with a description of components used on the test fixture). The instruction sheet includes values for the insertion loss of transmission lines, capacitors, and RF choke used in the test fixture, to simplify the calibration of test equipment.

The amplifier lineup includes models GAL-1 and GAL-21, the two broadest-frequency devices with coverage from DC to 8 GHz (and usable gain to 10 GHz). The former provides typical gain of 12.5 dB at 1 GHz and 11 dB at 8 GHz with gain flatness of 60.5 dB from DC to 2 GHz, while the latter offers typical gain of 13.9 dB at 1 GHz and 12.4 dB at 8 GHz with gain flatness of 60.6 dB from DC to 2 GHz. The GAL-1 achieves typical output power at 1-dB compression of +12.2 dBm, with typical noise figure of 4.5 dB and typical third-order intercept of +27 dBm (an effective dynamic range of 22.5 dB). The GAL-21 has typical output power at 1-dB compression of +12.6 dBm, with typical noise figure of 4 dB and typical third-order intercept of +27 dBm.

In terms of maximum output power, the models GAL-5 and GAL-6 provide typical output power at 1-dB compression of +18.0 and +18.2 dBm, respectively, with typical third-order intercepts of +35.0 and +35.5 dBm, respectively. The noise figures for the GAL-5 and GAL-6 amplifiers are 3.5 and 4.5 dB, respectively. Both models are designed for use from DC to 4 GHz, with gain of 16.0 and 11.3 dB, respectively, at 3 GHz.

The K1-GAL Designer's Kit offers a comprehensive, no-risk introduction to packaged monolithic HBT amplifiers. P&A: $99.95; stock. **Mini-Circuits, P.O. Box 350166, Brooklyn, NY 11235-0003; (718) 934-4500, FAX: (718) 332-4661, Internet: www.minicircuits.com.**

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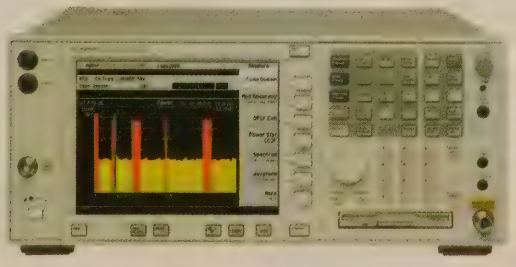
Spectrum Analyzers Simplify 3G Measurements

This series of spectrum analyzers makes a broad array of accurate RF power and digital demodulation measurements with one or two keystrokes.

JACK BROWNE
Publisher/Editor

Responsibility for meeting the signal-quality requirements of two-and-a-half-generation (2.5G) and third-generation (3G) wireless transmission systems has fallen to a high degree on amplifier and base-station designers. Their challenge is to design multicarrier amplifiers and other transmission-chain components that can maintain extremely high levels of linearity while accommodating more than a dozen signals, all of which are rapidly changing and placing huge loads on the transmitter (Tx). Measuring how well they perform under these conditions is a daunting task, even for veteran base-station designers. The measurements are complex, and the instruments designed to measure them have not been particularly easy to use. The PSA II series of spectrum analyzers from Agilent Technologies (Santa Rosa, CA) reduces the complexity of the measurement process, while retaining the ability to measure signal characteristics in excruciating detail when needed.



The PSA II series of spectrum analyzers features improved speed, accuracy, and frequency range compared to the first generation.

The PSA II is the second generation (2G) of the PSA series, introduced last year (see *Microwaves & RF*, November 2000, p. 137). The first-generation (1G) instrument covered 3.0 Hz to 26.5 GHz. The PSA II models **(see figure)** broaden the line to include coverage to 6.7 GHz (model E4443A) or 13.2 GHz (model E4445A). The instruments add capabilities for power and digital demodulation measurements that were missing from the first generation, and comprehensive phase-noise measurement capability. However, the new instruments' greatest contribution is to combine a simple, straightforward environment for wireless format-based RF power and digital demodulation measurements with a high-performance spectrum analyzer that can root out the cause of signal-transmission problems with evaluation of every important measurement parameter.

For example, a wideband-code-division-multiple-access (WCDMA) demodulation measurement can be made with three keystrokes, regardless of the instrument's configuration at the time. The user presses the "Mode" key, the

"WCDMA" key, and the desired measurement (such as code-domain power), and the PSA II performs the requested task to the specifications required by the appropriate air-interface standard. Measurements can be made in this way for any of six air-interface standards [WCDMA, Global System for Mobile Communications/Enhanced Data Rates for Global Evolution (GSM/EDGE), cdma2000, cdmaOne, North American Digital Communications (NADC), and personal digital cellular (PDC)]. All of these optional "measurement personalities" can simultaneously reside in the instrument.

If, however, a designer needs to truly dissect a signal, the PSA II provides an array of tools for spectrum analysis, power measurement, and digital demodulation. Power measurements include channel power, occupied bandwidth, multioffset adjacent-channel power, multicarrier power, power statistics (CCDF), harmonic distortion, burst power, third-order intercept, and spurious emissions with one-button setups for IS-95, J-STD-008, NADC, GSM/EDGE, WCDMA (3GPP), cdma2000, PDC, and Bluetooth.

The optional phase-noise measurement capability of the PSA II has been enhanced to support log plotting of phase noise over six decades of offset frequency, jitter measurements, as well as simultaneous display of instrument noise floor and phase noise.

Performance of the PSA II has also been improved. Swept measurements can be made at a rate up to 45 measurements/s with a 1-ms sweep time, which increases throughput in production, and provides immediate feedback from adjustments to the device under test (DUT). The measurement accuracy of the PSA II is typically ±0.21 dB to 3 GHz. A typical third-order intercept point (IP3) of +19 dBm translates into a dynamic range of greater than 116 dB at 1 GHz when aided by the 2-dB step attenuator.

The WCDMA adjacent-channel-power dynamic range is typically 72 dB (76 dB with noise correction), and phase noise at 1 GHz is typically –114 dBc/Hz at a 10-kHz offset, and better than –156 dBc/Hz at a 10-MHz offset. Sensitivity is high, typically –156 dBm at 1 GHz and as good as –170 dBm when the optional internal preamplifier is used. The PSA II offers 160 resolution bandwidths in 10-percent steps from 1 Hz to 3 MHz and fixed at 4, 5, 6, and 8 MHz for digitally swept or Fast Fourier transform (FFT) bandwidths. P&A: $42,000 and up; stock to 10 wks. **Agilent Technologies, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; (800) 452-4844 ext. 7554, Internet: www.agilent.com/find/psa.** MRF

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DESIGN

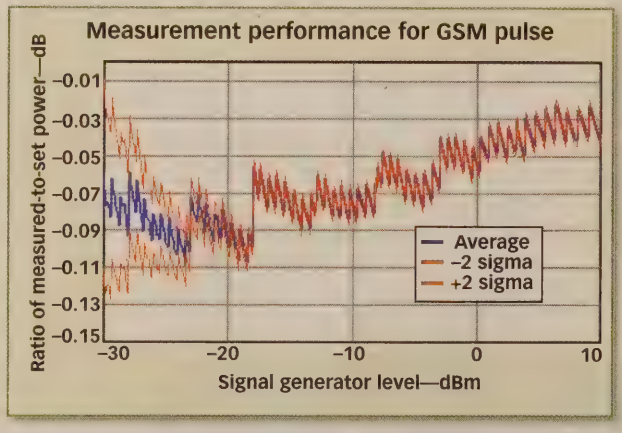
Continued from page 104

after them can also be captured with confidence.

Measurement throughput is an important factor too. Although the investigation of artifacts or special circumstances is important to the research-and-development (R&D) engineer, speed is king to the manufacturing engineer. In order to meet this need, the acquisition can be tuned to maximize throughput. This can be performed by continuous bandwidth-limited decimation if flat measurement bandwidths are important (i.e., when measuring peak power or peak-to-average ratio).

When signal-related triggering is not required, free-run acquisition can be used. In this mode, the acquisition simply takes fast samples of the power envelope and calculates the required measurement. It is also possible to randomize the position of sampling in this mode to increase signal coverage without reducing throughput.

This section presents an idea for measurements available from a sampling power meter. A typical application would be the power calibration of a GSM device. For this application, it would be normal to derive the trigger from the time-division-multiple-access (TDMA) pulse's rising edge and measure the average power within the useful part of the burst, avoiding overshoot at the start of the pulse. **Figure 9** shows how this measurement can be set up. The markers here define the start and stop times of a gated average-power measurement. **Figure 10** shows this measurement performed on an Agilent Technologies ESG-D signal generator using the EPM-P power meter and an E9320-series power sensor from the same company. Note that there is only ±1 percent (±0.04 dB) of measurement variation at −30 dBm. MRF



10. This graph shows the GSM power measurement on an Agilent Technologies ESG-D signal generator.

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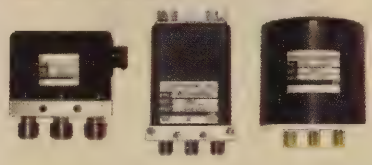

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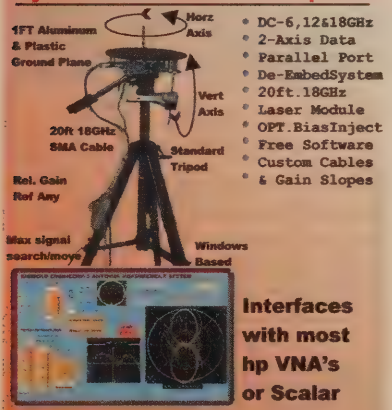

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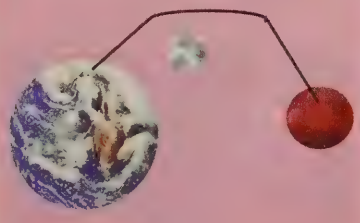

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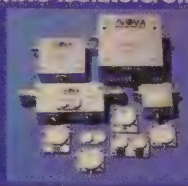

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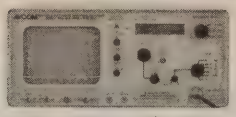

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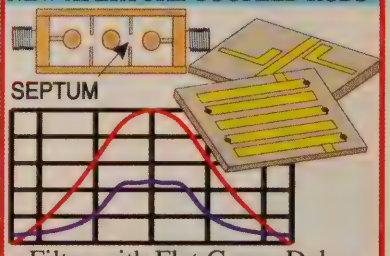

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trol as comparison, it can be extended to include lowpass filters, the antenna switch, etc. As long as the comparison is based on equivalent system functionality, the definition is valid. To understand effective efficiency mathematically, Eq. 16 can be rewritten as:

(See Eq. 23 below.)

Multiplying with η_{eff} in the numerator and denominator and rearranging the equation on both sides yields:

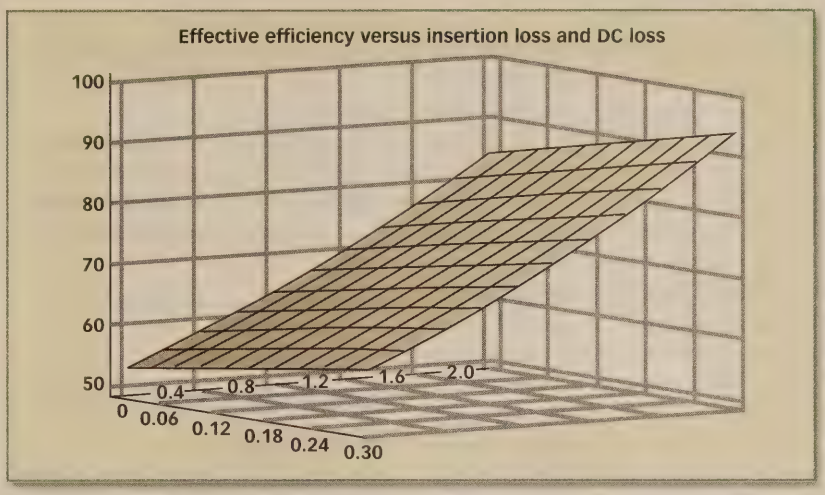
(See Eq. 24 below.)

η_{eff} is isolated and using the relationship that $(P_{out} + P_{loss})/\eta_{eff}$ equals delivered power to the PA, provides:

(See Eq. 25 below.)

After further algebraic manipulation, a second-order equation is seen:

(See Eq. 26 below.)



10. This is the graph of effective efficiency versus DC loss and insertion loss. The x-axis is ohmic loss, the y-axis is insertion loss from 0 to 2 dB, and the z-axis is effective efficiency in percent.

Here, P_{loss} and R are swept from 0 to 2 dB and from 0 to 0.3 Ω, respectively. η_{eff} is plotted in **Fig. 10.**

For example, a PA module with integrated power control has 54-percent η_{sys} at P_{out} = +35 dBm. Another PA module is placed in a discrete closed-loop system that senses power by using a directional coupler. The insertion loss in the directional coupler and stripline is assumed to be 0.5 dB. What would η_{eff} have to be to obtain 54-percent system efficiency? From Fig. 10, η_{eff} has to be 60.6 percent. In the previous case, the two modules were assumed to have no DC loss. If 200-mΩ DC loss is added η_{sys} for the PA with integrated power control that is reduced to 50.6 percent, whereas the required η_{eff} for the PA that is placed in a discrete power control is 61.5 percent. That is, the reference for the two PAs is exactly the same but the requirement to η_{eff} increases dramatically when small losses are added. As seen in Fig. 10, a 1-dB insertion loss and 200-mΩ DC loss requires an η_{eff} of approximately 73 percent. This is an important observation, which is often neglected when comparing PAs, and the end result is often that the total η_{sys} turns out to be less than expected.

After considering system implications of Tx design, it is clear that PAE does not directly relate to power consumption. On the contrary, a PA with good PAE can have higher power consumption over a PA with lower efficiency. Great efforts should be taken to eliminate supply losses and insertion losses. The easiest way to reduce power consumption is through system architecture and proper implementation rather than PA design. Based upon various power-control architectures, the most efficient architecture seems to be the supply-side method. Moreover, improvements can be made in system efficiency by optimizing harmonic levels in the PA and reducing the harmonic attenuation that is required in the lowpass filter. This may eliminate a lowpass filter, or at least reduce loss in the filter and improve total system efficiency.

Current levels of integration are changing the functional boundaries of PAs. Therefore, it is important to no longer discuss efficiencies in terms of PAE, but in the context of effective efficiency. MRF

$$\eta_{sys} = \frac{P_{pa} - P_{loss}}{\frac{P_{pa}}{\eta_{eff}} + P_r} \qquad \eta_{sys} = \frac{P_{out}}{\frac{P_{out} + P_{loss}}{\eta_{eff}} + P_r} \tag{23}$$

$$\eta_{eff}\left(P_r\eta_{sys} - P_{out}\right) + \eta_{sys}\left(P_{out} + P_{loss}\right) = 0 \tag{24}$$

$$\eta_{eff} = \eta_{sys}\left(\frac{P_{out} + P_{loss}}{P_{out} - {I_{dc}}^2 R\eta_{sys}}\right) \qquad \eta_{eff} = \eta_{sys}\left[\frac{P_{out} + P_{loss}}{P_{out} - \left(P_{out} + P_{loss}\right)^2\left(\eta_{eff}V_{dc}\right)^{-2}R\eta_{sys}}\right] \tag{25}$$

$$\left[P_{out}\left(\eta_{eff}\right)^2 - \eta_{eff}\eta_{sys}\left(P_{out} + P_{loss}\right) - \left(\frac{P_{out} + P_{loss}}{V_{dc}}\right)^2 R\eta_{sys}\right] = 0 \tag{26}$$

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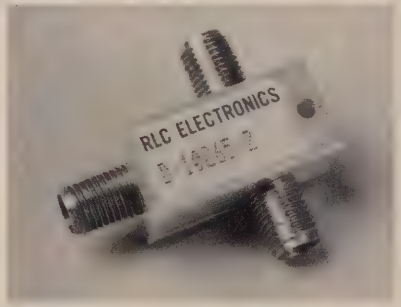

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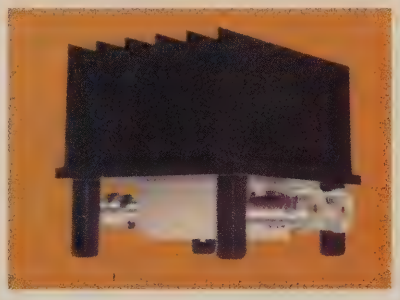

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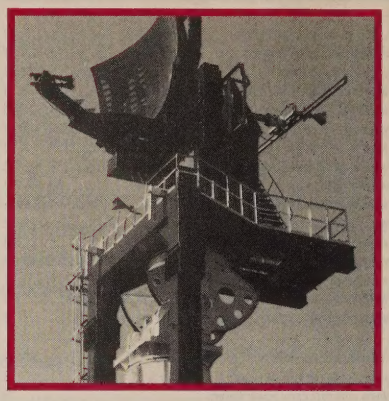
looking back



FOURTEEN YEARS AGO, a news story reported that Westinghouse (Baltimore, MD) had delivered its first solid-state ASR-9 airport radar (to replace a vacuum-tube version) as part of the Federal Aviation Administration's (FAA's) National Airspace System Plan (NASP) to overhaul the US air-traffic-control system.

next month

Microwaves & RF November Editorial Preview
Issue Theme: Military Electronics

News

High-power generation and amplification has long been a part of military systems design, and electron tubes have been the mainstay of high-power applications since their origins during World War II. For some uses, vacuum-tube devices may always be a more efficient and effective way of delivering power than solid-state devices, and current research aims not only to extend the efficiency of tubes, but to also expand their output-power capabilities, frequency ranges, and bandwidths. In a Special Report on high-power vacuum-tube devices, the November issue of *Microwaves & RF* will examine the latest tube technologies and performance levels.

Design Features

November's Design Features section offers a blend of articles on design and testing. For example, an article from Agilent Technologies' Communications Division presents techniques for improving the accuracy and speed of power measurements, notably in high-volume production environments. Additional articles continue a look at the meaning of power efficiency and how to achieve it in IC Tx designs. The second installment of an article series examines the design of LNA ICs, while the second half of an article describes methods for making triggered measurements with a sampling power meter.

Product Technology

The November Product Technology section will show that the benefits of direct frequency synthesis can indeed be delivered to the masses. Do not miss this exclusive review of a new, affordable, microwave signal generator that is based on direct synthesis, with submicrosecond switching speed and low phase noise for high-volume wireless testing. Additional product features will examine a low-power GPS Rx module that measures only 13 × 15 mm and a line of IC amplifiers for wireless applications.

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The "no-nonsense" attenuator...

For Audio, IF, and VHF.
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- SMA connectors, others available
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- Specs that beat the competition's

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Freq Range (MHz)	Atten Range (dB)	Atten vs Freq (dB)	Model No.
DC-60	40	±1.0	0682-40F
DC-100	15	±0.3	0682-15F
DC-100	30	±0.5	0682-30F
DC-250	10	±0.5	0682-10F
Uncalibrated models			
DC-60	40	±1.0	0682-40
DC-100	20	±0.6	0682-20
DC-100	30	±0.5	0682-30
DC-200	30	±2.0	0682-30A
DC-250	15	±1.2	0682-15
DC-500	10	±0.25	0682-10

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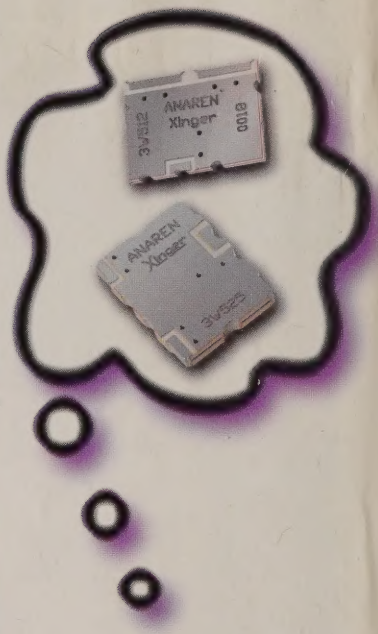
I'm thinking of an off-the-wall amplifier for commercial airline in-flight modems.



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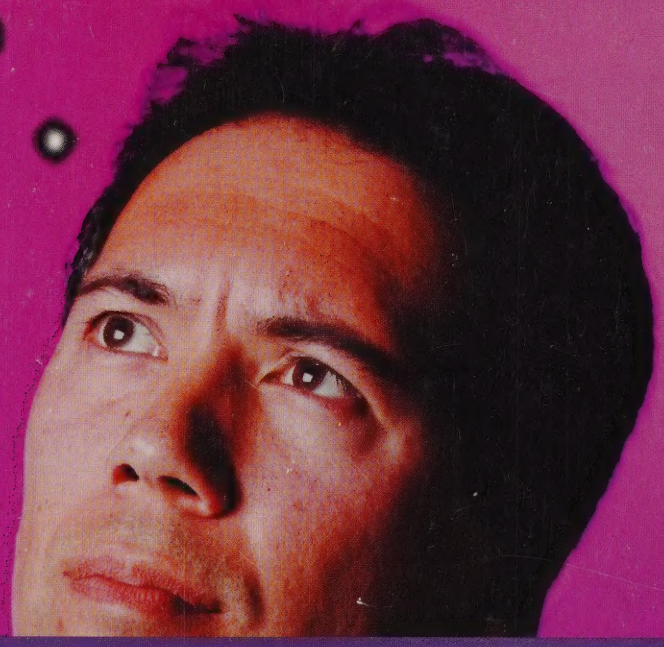

Then choose the microwave component lineup that'll help your idea fly.

Source at Anaren®, and you'll feel like a kid in a candy store. From Xinger®-brand couplers, power dividers, and baluns. To RF Power-brand resistors, terminations, and attenuators. To our full line of Ferrite isolators and circulators ... and more. Our fast growing, multi-format line of components will match your frequency. Performance specs. Manufacturing needs. And cycle times. We can even show you cost- and space-saving ways to integrate them. Or devise a customized version of an off-the-shelf part — in four weeks or less. Take the first step to being our components partner: get your free Anaren "Thinking Kit." To reply, use the reader service number.

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